Modeling and Prediction of Electromagnetic Compatibility (EMC) Performance for GPS Systems in Mixed-Signal and RF Environments

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Abstract

Global Navigation Satellite System receivers operate in crowded electromagnetic conditions where small, unintended emissions and strong in-band or adjacent-band signals can alter timing and positioning observables. In mixed-signal architectures that combine sensitive RF front-ends, high-resolution data converters, and dense digital logic, electromagnetic compatibility challenges arise from distributed coupling paths that span packages, interconnects, and enclosures. The resulting interference does not behave as a single deterministic blemish; rather, it manifests as a variable environment whose statistics, modal structure, and nonlinearity depend on layout decisions, clocking strategies, and front-end bias points. This paper investigates a modeling and prediction framework for electromagnetic compatibility performance of GPS systems in mixed-signal and RF environments, emphasizing transparent abstractions that connect field-level descriptions, circuit macromodels, and system metrics. The approach integrates multiport representations of packages and printed-circuit boards with stochastic field descriptions, enabling consistent propagation of uncertainty from emissions through coupling to desensitization metrics at the correlator. The analysis explores the role of nonlinearity in low-noise amplifiers and mixers, converter aperture jitter and quantization noise under interference, and power integrity fluctuations that translate conducted noise into phase and gain perturbations. Attention is given to statistical indicators of robustness that align with compliance practices while avoiding overreliance on worst-case assumptions. The narrative maintains a neutral tone, focusing on reproducible formulations, algorithmic steps for prediction, and validation procedures that connect bench measurements with over-the-air results. The outcome is a cohesive set of models suitable for early-stage design exploration and for later diagnostic use when empirical artifacts emerge.

1 Introduction

Positioning systems that employ spread-spectrum signaling derive their robustness from the deliberate dilution of spectral energy across a wide bandwidth, enabling operation even in the presence of significant interference and multipath distortion [1]. The fundamental premise is that by spreading signal energy uniformly over frequency, the receiver can recover information through correlation processing while treating most narrowband interferers as noise. Yet, the practical resilience of such systems depends not only on this spreading gain but also on the complex interplay between the electromagnetic environment, hardware architecture, and nonlinearities within the receiver chain. Electromagnetic coupling paths, parasitic impedances, and imperfect isolation between analog and digital subsystems can significantly alter how external energy is received and processed, leading to distortion, cross-modulation, or unintentional amplification of interference components. Mixed-signal implementations are particularly vulnerable to these effects, as they co-locate sensitive high-gain analog circuits with fast-switching digital logic on densely integrated substrates [2]. The resulting electromagnetic compatibility (EMC) challenges introduce variability that cannot be captured by a single performance figure; rather, receiver robustness must be viewed as a statistical distribution across operating conditions, manufacturing tolerances, and environmental contexts.

Parameter	**Symbol**	**Unit**	**Nominal**	**Range**
Carrier-to-noise density ratio	C/N_0	dB-Hz	45	3550
Loop bandwidth	B_L	Hz	15	525
Code chip rate	R_c	Mcps	1.023	0.52.0
Sampling rate	f_s	MHz	20	1040

Table 1. Receiver signal and tracking parameters.

A system-level understanding begins with the abstraction of physical fields and currents into a finite-dimensional network representation that preserves the essential coupling mechanisms while enabling tractable computation. This mapping translates the continuous electromagnetic domain into a circuit-theoretic framework, where voltages and currents at ports or nodes represent aggregated field interactions. The resulting network model can incorporate both distributed coupling effects and localized nonlinearities through multiport impedances and behavioral elements. Once in this domain, standard linear-algebraic toolsmatrix factorization, eigenvalue analysis, and stochastic propagationcan be used to quantify how interference energy couples through various paths into the receivers signal chain [3]. This reduction provides a bridge between physical layout phenomena and higher-level performance metrics relevant to navigation systems.

From this network representation, one can trace the influence of interference through the receivers spectral processing stages, where despreading and tracking loops translate physical signals into timing and carrier observables. In global navigation satellite systems (GNSS) such as GPS, these observables include code phase and carrier phase, which together define the time-of-arrival and Doppler measurements used for position estimation. Interference manifests in these domains by reducing the ratio of despread carrier power to effective noise densityoften summarized as the carrier-to-noise-density ratio, $C/N_0$and by distorting discriminator characteristics that guide loop convergence [4]. Specifically, amplitude modulation, phase noise, and spectral asymmetry in the interference can bias code and carrier tracking discriminators, leading to systematic timing errors and degraded stability. Since the aggressor spectrum is rarely fixed or deterministic, performance must be evaluated statistically over ensembles of interference scenarios, bandwidth allocations, and dynamic input levels. Such probabilistic characterization captures both average degradation and tail-risk behavior, identifying conditions under which tracking failure or cycle slips become probable.

To predict these effects in a realistic and computationally feasible manner, a multi-layer modeling approach is advantageous. At the outer layer, the ambient electromagnetic environment is represented stochastically, using random processes or distributions to describe power spectral densities, polarization states, and temporal dynamics of interference sources [5]. This layer captures environmental variability arising from co-located transmitters, reflections, and dynamic loading of shared spectrum. The intermediate layer consists of multiport circuit macromodels describing the printed-circuit board, interconnects, and packaging structures. These models incorporate coupling paths, parasitic elements, and resonance behaviors derived from full-wave simulations or measurements, yet are compact enough to support system-level statistical analysis. At the innermost layer, the analog front-end is described by nonlinear behavioral modelssuch as Volterra or X-parameter representationsthat capture saturation, gain compression, and harmonic distortion [6]. The digital stages, including mixers and correlators, are treated as deterministic transformations conditioned on the analog input but parameterized by quantization effects and clock jitter.

Combining these layers yields a joint probabilistic model of the receivers behavior under interference. By sampling from or analytically propagating distributions through this composite model, one can perform uncertainty quantification to estimate how likely key performance indicators (KPIs) are to degrade below mission thresholds. These KPIs may include C/N_0, tracking jitter, acquisition probability, and navigation accuracy. Monte Carlo methods or polynomial chaos expansions can approximate the forward propagation of uncertainty, providing not just mean estimates but full probability distributions of performance metrics [7]. This enables system designers to assess not merely worst-case margins but confidence intervals, quantiles, and conditional probabilities that reflect true operational diversity.

Model Layer	Representation	Key Variables	Nonlinearity	Stochastic Inputs	Output Metrics
Environment	Spectral process	PSD, polarization	–	✓	Coupling fields
Circuit board	Multiport network	Z_{ij}, Y_{ij}	–	✓	Port voltages
Front-end	Behavioral model	Gain, phase	✓	–	Distortion levels
Digital stage	Deterministic map	Quantization, timing	✓	✓	Tracking error

Table 2. Hierarchical modeling layers and associated attributes.

Beyond prediction, the same probabilistic machinery can be inverted for diagnostic and inference tasks. When measurements of degraded performance are availablesuch as unexpected C/N_0 drops, loop instability, or anomalous residualsBayesian inversion allows identification of likely root causes within the electromagnetic or circuit domain. By conditioning the joint model on observed data, posterior distributions over latent variables, such as coupling coefficients or nonlinear gain parameters, can be computed. This diagnostic capability facilitates targeted mitigationshielding redesign, grounding adjustments, or filter retuningwithout exhaustive experimentation.

The overall perspective treats the receiver not as a static deterministic entity but as a stochastic dynamical system embedded within an uncertain electromagnetic environment. Its robustness is inherently distributional: for any given configuration and context, performance outcomes are random variables whose statistics depend on both hardware characteristics and environmental fluctuations. This recognition aligns modeling, measurement, and inference within a unified probabilistic framework. It provides the tools needed to design, analyze, and validate spread-spectrum positioning systems that must maintain integrity in the face of dense spectral competition, hardware nonidealities, and pervasive environmental variability [8]. In doing so, it bridges physical electromagnetics, circuit theory, and statistical signal processing, creating a pathway toward predictive and diagnostic models that support both robust design and intelligent adaptation in next-generation navigation receivers.

Parameter	Symbol	Unit	Nominal	Range
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Table 3. Receiver signal and tracking parameters.

Source spectra are allowed to vary within envelopes, spatial correlation in fields is represented with parameterized kernels, and device nonlinearity is captured without presupposing a specific dominant term. Computations are organized so that intermediate quantities remain interpretable, thereby enabling design trade studies that compare routing options, enclosure modifications, and filtering choices under consistent metrics.

$$\text{Let } \mathcal{H}: \ \mathbf{e}(\mathbf{r}, t) \mapsto \mathbf{v}(t), \qquad \mathbf{v}(t) \in \mathbb{R}^p$$

2 EMC Fundamentals in Mixed-Signal GPS Receivers

Electromagnetic compatibility in this context concerns the coexistence of the desired satellite signals with spurious emissions, intentional interferers, and self-generated switching noise [9]. The GPS signal arrives at very low power relative to thermal noise, and the receiver design relies on coherent and noncoherent integration to accumulate processing gain. Any interference that couples prior to correlation can reduce the effective gain by saturating the front-end, increasing the apparent noise density, or rotating the signal constellation in a manner that degrades discriminator slopes. The multi-octave environment of a mixed-signal board introduces coupling through radiated and conducted paths, and the paths are modulated by layout geometries, ground references, and supply impedances.

A minimal field-to-circuit abstraction is obtained by viewing the assembly as a multiport where each port aggregates a region of current return, pad cluster, or enclosure seam. The external electromagnetic environment is summarized by equivalent Thevenin or Norton representations at those ports, while internal switching activity appears as distributed current sources [10]. The ports are linked by a frequency-dependent impedance matrix that captures mutual inductance, capacitance, and loss. In this description, the aggressor-to-victim transfer is encoded in off-diagonal blocks of the admittance or scattering matrices, and the sensitivity of the receiver is determined by how those transfers align with front-end selectivity and linearity.

$$\mathbf{i}(\omega) = \mathbf{Y}(\omega)\,\mathbf{v}(\omega), \qquad \begin{bmatrix} \mathbf{v}_{\text{vict}} \\ \mathbf{v}_{\text{agg}} \end{bmatrix} = \begin{bmatrix} \mathbf{Z}_{vv} & \mathbf{Z}_{va} \\ \mathbf{Z}_{av} & \mathbf{Z}_{aa} \end{bmatrix} \begin{bmatrix} \mathbf{i}_{\text{vict}} \\ \mathbf{i}_{\text{agg}} \end{bmatrix}$$

When digital edges switch, displacement current couples through parasitics into analog traces. The spectral signature is shaped by rise time, line impedance, and terminations, and it is further colored by enclosure modes. At RF, the effective cross-section of seams and apertures determines radiative leakage, while cable harnesses act as unintentional antennas that can inject common-mode currents [11]. Those currents convert to differential voltage at asymmetries and terminations, with the conversion factor determined by imbalance and return path discontinuities. In the presence of GPS band filters, the residual energy can still interact nonlinearly with active devices, creating intermodulation products that fall into the spread-spectrum processing bandwidth.

$$\mathbf{s}(\omega) = \mathbf{S}(\omega)\,\mathbf{a}(\omega), \qquad \|\mathbf{S}\|_2 \le 1, \quad \mathbf{S}^\dagger\mathbf{S} = \mathbf{I} \text{ on matched ports}$$

3 Interference Taxonomy and Coupling Pathways

Interference is categorized by origin and spectral structure. Out-of-band emitters create blockers that either compress gain or, through device nonlinearity, downconvert energy into the passband [12]. In-band but uncorrelated signals increase the noise floor after despreading. Bursty or cyclostationary sources such as switched-mode power stages produce sidebands whose alignment with the code and carrier loop bandwidths determines the observed jitter. Self-generated interference from digital logic exhibits spatial coherence patterns related to clock distribution and return paths, and its harmonics can intersect with local oscillator spurs to produce dense spectral line sets.

The coupling can be decomposed into differential and common-mode channels, with conversion operators mapping one to the other at discontinuities. The board-level power distribution network carries conducted noise that appears as supply modulation at the low-noise amplifier and mixer [13]. That modulation results in AM-to-PM conversion and in parametric variations of transconductance and bias points, which translate into gain and phase noise increments. In enclosures with finite conductivity seams, resonant modes couple to cable shields, and the shield transfer impedance defines the leakage into the inner conductors. The relative phase between leakage into the antenna port and leakage into intermediate stages governs constructive or destructive interference at the correlator output.

$$\begin{aligned}
\mathbf{v}_{\mathrm{IF}}(\omega) &= \mathbf{T}_{\mathrm{EM}\to\mathrm{IF}}(\omega)\,\mathbf{e}_{\mathrm{ext}}(\omega) + \mathbf{T}_{\mathrm{dig}\to\mathrm{IF}}(\omega)\,\mathbf{i}_{\mathrm{sw}}(\omega)\\
\mathbf{T}_{\mathrm{EM}\to\mathrm{IF}}(\omega) &= \mathbf{H}_{\mathrm{RF}}(\omega)\,\mathbf{C}_{\mathrm{rad}}(\omega)\,\mathbf{G}_{\mathrm{encl}}(\omega)\,\mathbf{C}_{\mathrm{pcb}}(\omega)
\end{aligned}$$

The antenna and preselect filter shape the first line of defense, but the filter skirts and group delay ripple influence how adjacent-band energy excites intermodulation pathways [14]. The mixer local oscillator leakage and the image response articulate further sensitivities. Code tracking loops integrate across chip periods to suppress spectral lines that do not align with the spreading sequence, yet residual narrowband components introduce bias in the discriminator when the in-phase and quadrature branches are unbalanced. Carrier tracking reacts to both phase and frequency variations caused by AM-to-PM conversion in the presence of blockers.

$$r[n] = \sum_k s[k]\,c[n-k] + w[n] + \iota[n], \qquad \widehat{P}_{\mathrm{despread}} = \frac{1}{N}\left|\sum_{n=0}^{N-1} r[n]\,c[n]\right|^2$$

4 FieldCircuit Co-Modeling and Stochastic EMI Environments

Field Model	Quantity	Type	Description	Symbol
Incident field	Random vector process	Complex Gaussian	External EMI excitation	$\mathbf{E}(\mathbf{r},\omega)$
Covariance kernel	Hermitian matrix	Second-order statistic	Spatialfrequency correlation	$\mathbf{K}(\mathbf{r},\mathbf{r}',\omega)$
Port mapping	Integral operator	Linear transformation	Field-to-network coupling	$\mathbf{\Psi}_p(\mathbf{r},\omega)$
Network response	Multiport impedance	Deterministic	Converts port forces to voltages/currents	$\mathbf{Z}(\omega)$

Table 4. Key components of the stochastic fieldcircuit co-model.

A consistent framework begins with a stochastic model for the external electromagnetic field that impinges on the enclosure and harness, combined with a network model for internal coupling. The field is treated as a zero-mean vector process with frequency-dependent covariance, and the enclosure is represented by modal transfer operators that map incident fields to port voltages and currents [15]. The network that represents the board and package is described by an impedance or scattering matrix that is extracted from simulation or measurement, and internal switching sources are represented by cyclostationary processes with known second-order statistics.

$$\mathbb{E}\{\mathbf{E}(\mathbf{r},\omega)\} = \mathbf{0}, \quad \mathbb{E}\{\mathbf{E}(\mathbf{r},\omega)\mathbf{E}^{\dagger}(\mathbf{r}',\omega')\} = \delta(\omega-\omega')\,\mathbf{K}(\mathbf{r},\mathbf{r}',\omega)$$

The mapping from fields to network ports is expressed with radiation and reception integrals that define equivalent electromotive forces at each port. Those forces drive the network together with internal sources, and the resulting port responses propagate to the RF front-end through the interconnect. The stochastic structure is preserved under linear transformations, permitting analytic computation of response covariances [16]. Nonlinear stages introduce higher-order statistics that require approximations such as Volterra series or generalized polynomial chaos expansions to propagate uncertainty.

$$\mathbf{v}_p(\omega) = \int_{\Gamma} \mathbf{\Psi}_p(\mathbf{r},\omega)\cdot\mathbf{E}(\mathbf{r},\omega)\,dS, \qquad \mathbf{v}(\omega) = \mathbf{Z}(\omega)\,\mathbf{i}(\omega) + \mathbf{v}_p(\omega) + \mathbf{v}_{\mathrm{int}}(\omega)$$

With these representations, the interference at the input of the low-noise amplifier is described by a complex Gaussian process whose spectrum depends on both external statistics and internal coupling. The processing gain of the spread-spectrum receiver shapes this spectrum through correlation with the known codes. A central quantity for prediction is the distribution of the despread power over integration intervals, which can be obtained by integrating the response covariance over the code kernel. When the external field exhibits partial coherence, the covariance kernel is not diagonal in frequency, and the effective interference after despreading reflects that cross-correlation. [17]

$$\mathbb{E}\{\widehat{P}_{\mathrm{despread}}\} = \mathbf{h}^{\dagger}\mathbf{R}_{vv}\mathbf{h}, \qquad \mathrm{Var}\{\widehat{P}_{\mathrm{despread}}\} = 2\,\mathbf{h}^{\dagger}\mathbf{R}_{vv}\mathbf{h}\,\mathbf{h}^{\dagger}\mathbf{R}_{vv}\mathbf{h}$$

5 Nonlinear RF Front-End Dynamics and Desensitization Metrics

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Nonlinearity Source	Model	Effect	Metric	Expression
LNA cubic term	$h_3(\tau_1,\tau_2,\tau_3)$	Intermodulation	IM3 power	P_{IM3}
Mixer conversion loss	Amplitude-dependent	Desensitization	$\Delta C/N_0$	Eq. (25)
ADC aperture jitter	Timing noise	Baseband noise floor	$S_{n,\mathrm{jitter}}$	$(\omega\sigma_t)^2 S_x$
Quantization	Uniform quantizer	Spectral shaping	$S_q(\omega)$	$\frac{\Delta^2}{12}\lvert H_d\rvert^2$

Table 5. Principal nonlinear front-end mechanisms and their analytic indicators.

The low-noise amplifier and mixer are modeled as weakly nonlinear systems whose behavior under interference determines desensitization. A cubic polynomial captures third-order intermodulation when blockers are strong yet not in deep compression, and a memory kernel accounts for frequency-dependent gain and phase. The cubic term produces combinations that can fall into the GPS processing bandwidth even when blockers are well outside the passband, and the resulting intermodulation increases the in-band noise density. The onset of compression reduces the effective carrier gain and flattens discriminator slopes, further contributing to tracking noise. [18]

$$y(t) = \int h_1(\tau)\, x(t-\tau)\, d\tau + \iint h_2(\tau_1, \tau_2)\, x(t-\tau_1)x(t-\tau_2)\, d\tau_1 d\tau_2 + \iiint h_3(\tau_1, \tau_2, \tau_3) \prod_{k=1}^{3} x(t-\tau_k)\, d\tau_1 d\tau_2 d\tau_3$$

Under a two-tone blocker with angular frequencies that straddle the passband, third-order mixing creates an in-band component whose amplitude scales with the cube of the blocker amplitude. When multiple blockers exist, cross terms appear that require the full Volterra formulation to predict. The noise figure under interference becomes a function of the blocker amplitudes and phases, and the effective carrier-to-noise density ratio at the correlator reflects both gain compression and added intermodulation noise. The acquisition and tracking loops respond to these changes by adjusting loop bandwidths and integration times, which in turn modifies the system sensitivity to cyclostationary interference. [19]

$$P_{\mathrm{IM3}}(\omega_0) \approx \frac{3}{4} \left|H_3(2\omega_1 - \omega_2, \omega_1, \omega_1)\right|^2 P_1^2 P_2 \quad \Rightarrow \quad \Delta C/N_0 \simeq -10 \log_{10}\left(1 + \frac{P_{\mathrm{IM3}}}{N_0 B}\right)$$

The mixer contributes conversion loss variations that depend on local oscillator amplitude and on port impedances as seen through the coupling network. Local oscillator leakage interacts with front-end nonlinearity to generate spurs that may align with code harmonics, and the spur set broadens under supply modulation. Aperture jitter in the analog-to-digital converter converts high-frequency interference into baseband noise proportional to signal slope, and quantization in the presence of interfering tones produces non-uniform error spectra [20]. These effects require a joint evaluation in which the interference spectrum at the converter input, the converter timing noise, and the digital decimation filters are represented within a single analysis.

$$S_{n,\mathrm{jitter}}(\omega) \approx (\omega \sigma_t)^2\, S_x(\omega), \qquad S_q(\omega) = \frac{\Delta^2}{12} \cdot \left|H_{\mathrm{d}}(\omega)\right|^2$$

6 PCB, Package, and Power Integrity Effects

The printed-circuit board and package define the parasitic environment that mediates coupling and determines how conducted noise appears as RF impairments. Power distribution networks with frequency-dependent impedance translate current transients into supply voltage fluctuations, and those fluctuations modulate the front-end. The multiport description of the power network provides a route to compute the transfer from digital load currents to sensitive analog rails, and the same network reveals how return path discontinuities convert common-mode to differential-mode voltages on signal pairs. The geometric arrangement of planes, stitching vias, and slots sets modal resonances that can align with clock harmonics. [21]

$$\mathbf{v}_{\mathrm{rail}}(\omega) = \mathbf{Z}_{\mathrm{PDN}}(\omega)\, \mathbf{i}_{\mathrm{load}}(\omega), \qquad \Delta\phi(\omega) \approx k_\phi(\omega)\, v_{\mathrm{rail}}(\omega), \quad \Delta g(\omega) \approx k_g(\omega)\, v_{\mathrm{rail}}(\omega)$$

Package inductances and capacitances determine the effective isolation between the antenna input and digital pins. Even small mutual inductance values introduce measurable coupling at RF, and the effective isolation varies with frequency according to the package and board resonances. When cables are attached, the shield transfer impedance and termination quality dictate how external fields convert to inner conductor noise, and the board-level common-mode chokes provide only finite attenuation subject to saturation and frequency-dependent parasitics. The presence of active bias networks adds a further path because bias tees and current mirrors respond to conducted interference and relay it to gain stages.

$$\begin{bmatrix}\mathbf{v}_{\text{ana}}\\ \mathbf{v}_{\text{dig}}\end{bmatrix} = \begin{bmatrix}\mathbf{Z}_{aa} & \mathbf{Z}_{ad}\\ \mathbf{Z}_{da} & \mathbf{Z}_{dd}\end{bmatrix}\begin{bmatrix}\mathbf{i}_{\text{ana}}\\ \mathbf{i}_{\text{dig}}\end{bmatrix}, \qquad \mathbf{Z}_{ad}(\omega) \neq \mathbf{0} \Rightarrow \text{coupling}$$

The power integrity design must therefore be considered jointly with RF filtering and routing [22]. Decoupling strategies that reduce the impedance magnitude at switching harmonics shrink the amplitude of AM-to-PM conversion at the LNA and mixer, while careful via placement and return stitching lower common-mode conversion. Enclosure seams and apertures are likewise integrated into the model so that radiated and conducted mechanisms are treated within a single transfer description.

$$|Z_{\text{PDN}}(\omega)| \approx \left| \left(j\omega L_{\text{pkg}} + \sum_k \frac{1}{j\omega C_k} \parallel j\omega L_k + R_k \right) + Z_{\text{plane}}(\omega) \right|$$

7 Prediction Framework and Uncertainty Quantification

Prediction proceeds by constructing stochastic inputs and propagating them through linear and nonlinear blocks to obtain distributions for system metrics. The external field is parameterized with a small number of hyperparameters that govern spectral shape and spatial coherence, and the internal switching sources are specified by rise times, edge rates, and activity factors [23]. The network blocks are frequency-dependent matrices measured or simulated, and the nonlinear front-end is represented with kernels or with polynomial terms fitted to device data. Posterior distributions for the metrics are obtained by Monte Carlo or by spectral expansions that exploit polynomial chaos bases in the space of the random inputs.

Coupling Path	Representation	Dominant Elements	Frequency Dependence	Impact
Power rail	$\mathbf{Z}_{\text{PDN}}(\omega)$	Decaps, vias, planes	Resonant	AMPM conversion
Analogdigital cross-talk	$\mathbf{Z}_{ad}(\omega)$	Package pins, bondwires	Broadband	Desensitization
Cable shield	Transfer impedance	Braiding, termination	Linear rise	Common-mode injection
Enclosure seam	Slot admittance	Gaps, screws, apertures	Resonant	Radiated leakage

Table 6. Representative PCB and package coupling mechanisms relevant to RF integrity.

$$\mathcal{M} : \boldsymbol{\theta} \mapsto \mathbf{y}, \quad \boldsymbol{\theta} \sim p(\boldsymbol{\theta}), \quad \mathbf{y} = \Phi(\boldsymbol{\theta}) = \text{Metrics}\left(\mathbf{T}(\boldsymbol{\theta}), \mathbf{H}_{\text{NL}}(\boldsymbol{\theta}), \mathbf{S}_{\text{EMI}}(\boldsymbol{\theta})\right)$$

Key GPS-oriented metrics include the distribution of carrier-to-noise density at the correlator after integration, the probability that tracking loops lose lock under given interference statistics, and the dispersion of code and carrier jitter. The code discriminator is modeled with its slope and noise-equivalent bandwidth, and the carrier loop is described with a linearized phase model. Under interference, the noise terms include thermal contributions, intermodulation noise from the LNA and mixer, and jitter-induced noise from the converter [24]. The combined noise drives the loop error dynamics, and the lock conditions translate to bounds on the phase variance. Uncertainty propagation yields probabilities of surpassing those bounds.

$$\dot{\phi}(t) = \omega_\Delta(t) + n_\phi(t), \qquad \sigma_\phi^2 = \int_0^\infty |H_{\text{PLL}}(j\omega)|^2 \, S_{n_\phi}(\omega) \, \frac{d\omega}{2\pi}$$

Because the environment is variable, prediction emphasizes quantiles of performance rather than single-point values. Percentile curves for carrier-to-noise density and for code jitter across interference hyperparameters provide a more informative picture than worst-case margins alone [25]. The framework admits sensitivity analysis by computing Sobol indices or local derivatives with respect to inputs, which helps identify which coupling matrices or nonlinear coefficients exert the largest influence on system outcomes within the modeled scenarios.

$$\mathrm{Var}(Y) = \sum_i V_i + \sum_{i<j} V_{ij} + \cdots, \qquad S_i = \frac{V_i}{\mathrm{Var}(Y)}, \quad S_{ij} = \frac{V_{ij}}{\mathrm{Var}(Y)}$$

8 Carrier-to-Noise Density as a Governing Metric for GPS Tracking Robustness

Carrier-to-noise density is a compact descriptor for how much useful signal energy survives at the correlator relative to the spectral density of noise and interference at the same point in the chain. The quantity is naturally expressed in linear units as a ratio of watts to watts per hertz and in practice is reported in dB-Hz. Because correlation aggregates energy over a finite dwell, the ratio determines the effective signal-to-noise ratio after an integration interval and thereby controls the slope and variance of the code and carrier discriminators. In mixed-signal receivers operating amid emissions and blockers, the value is not a static number; it depends on thermal noise set by the system noise temperature, on interference that raises the effective density beyond thermal levels, and on nonidealities that scatter or compress the signal [26]. A working reliability target near 35 dB-Hz is commonly employed to ensure resilient tracking under benign dynamics, and excursions below that level tend to increase jitter and cycle slips. The function of modeling is to translate front-end conditions into predictions for the ratio and to quantify how often and by how much it will be depressed in representative environments.

A minimal signal model at the correlator input writes the complex baseband sample stream as a superposition of a spread-spectrum component, additive noise with approximately white spectrum over the loop bandwidths of interest, and interference with arbitrary structure. The despreading operation multiplies by the code replica and integrates over a dwell, so the post-correlation statistic consists of a coherent contribution proportional to the carrier power and a noise-plus-interference contribution whose variance scales with the noise spectral density times the dwell [27]. In linear units, the post-correlation coherent integration signal-to-noise ratio obeys

$$\mathrm{SNR}_T = \left(\frac{C}{N_0}\right) T,$$

where T is the coherent dwell. When noncoherent combining is used across L dwells, the effective ratio increases further only sublinearly because the phase information is discarded; a common approximation invokes a square-root growth for detection statistics, and refinements use the noncentral chi-square family to represent the distribution of the envelope. The coherent relation underscores why carrier-to-noise density is the natural predictor for both acquisition and tracking performance and why any mechanism that reduces the effective ratio manifests as degraded stability of the loops.

At RF and intermediate frequency, the signal power available to the correlator is determined by the link budget and by the analog chain [28]. Denoting the received carrier power at the antenna terminals by C_{ant} and the system noise temperature by T_{sys}, the linear ratio at the correlator under idealized filtering can be related to

$$\left(\frac{C}{N_0}\right)_{\text{lin}} \approx \frac{G_{\text{RF}}\,\eta_{\text{corr}}\,C_{\text{ant}}}{k\,T_{\text{sys}}},$$

where G_{RF} is the cascaded power gain from the antenna terminals to the correlator, η_{corr} represents correlator and matched filtering losses, and k is Boltzmanns constant. The term T_{sys} aggregates antenna sky temperature, amplifier noise contributions, and excess temperature from losses ahead of the low-noise amplifier. In mixed-signal receivers, T_{sys} is not strictly a temperature; interference and switching noise increment the effective density by an amount N_0^{int} so that the denominator becomes $kT_{\text{sys}} + N_0^{\text{int}}$. This raises the importance of electromagnetic compatibility because any conducted or radiated coupling that injects energy within or near the processing bandwidth translates directly to a reduction of the ratio.

Carrier tracking error can be approximated by a linearized phase model with an equivalent noise term that encapsulates thermal, interference, and implementation noise. With a one-sided loop noise bandwidth B_n for a second-order or third-order loop operating in small-error regime, a widely used approximation connects phase jitter variance to the ratio via

$$\sigma_\phi^2 \approx \frac{B_n}{C/N_0},$$

with σ_ϕ in radians [29]. The constant of proportionality depends on loop order, discriminator, and gain normalization, yet the scaling with the inverse of the ratio persists across implementations. Therefore, if the ratio drops due to interference or gain compression, the carrier phase jitter spreads and cycle slips become more likely. Frequency tracking exhibits a similar relationship with an additional factor accounting for the loop structure, and the dynamic stress errorarising from unmodeled jerk and accelerationadds in quadrature with the noise-driven jitter so that a margin above the threshold is necessary for mobile or high-dynamics scenarios.

Code tracking depends on the slope of the early-late discriminator, the correlation loss due to data bit transitions, and the noise level at the prompt outputs. If S_τ denotes the magnitude of the derivative of the correlator output with respect to delay at the operating point, a nominal variance expression for the code delay error under white noise gives [30]

$$\sigma_\tau^2 \approx \frac{N_0}{2C}\,\frac{B_{\text{DLL}}}{S_\tau^2},$$

where B_{DLL} is the code loop noise bandwidth. Introducing $\text{SNR}_T = (C/N_0)T$ and expressing S_τ in terms of code chip rate and early-late spacing recasts the relation into a form that again shows $\sigma_\tau \propto (C/N_0)^{-1/2}$. The proportionality invites the same conclusion as for the carrier: when the ratio decreases, the code jitter increases, and if the variance grows beyond the capture range of the loop under its current bandwidth, code lock is lost. The interaction with data bit transitions and with pilot/data channel configurations then becomes important because bit-edges effectively limit coherent dwell and thereby cap the usable SNR_T for acquisition and tracking.

The ratio reported by a receiver is typically estimated from prompt-power and noise-power measurements over a known bandwidth. In a quadrature architecture with automatic gain control, the estimator often leverages the variance of the in-phase and quadrature samples away from the correlation peaks to infer the noise density and compares that to the prompt channel energy. If P_{prompt} denotes the mean squared magnitude at the prompt output after despreading and N denotes an estimate of the in-band noise power over the same bandwidth, a textbook estimator in linear units forms

$$\widehat{\left(\frac{C}{N_0}\right)} = \frac{P_{\text{prompt}} - N}{N} \frac{1}{T},$$

with bias corrections applied for finite-sample effects and for squaring loss in envelope-based measurements [31]. In devices with additional whitening or interference mitigation blocks, the estimator must be calibrated to the modified effective bandwidth and to any spectral notches that alter the relationship between the measured noise power and the density seen by the loop. The calibration becomes especially relevant when interference mitigation is active because notches that excise narrowband components can change the shape of the noise power spectral density in ways that are not captured by a scalar density value.

Interference modifies the ratio through multiple mechanisms. A narrowband blocker at an offset from the carrier frequency contributes a line in the baseband spectrum after mixing, and if the blocker is strong, the nonlinearity in the low-noise amplifier and mixer produces intermodulation that spreads energy into the passband. The upshot is an increment to the effective N_0 seen by the correlator [32]. Let $S_i(\omega)$ denote the interference spectrum referred to the correlator input; the increment to the in-band density is then

$$N_0^{\text{int}} = \frac{1}{B_{\text{eff}}} \int_{-B_{\text{eff}}/2}^{B_{\text{eff}}/2} S_i(\omega)\, d\omega,$$

where B_{eff} is the effective equivalent noise bandwidth of the front-end and digital filtering. Nonstationary or cyclostationary interference requires time-frequency treatment, and the density becomes a function of time so that the ratio is best considered as a stochastic process. In that view, reliability statements adopt a probabilistic form, such as the fraction of time the ratio exceeds a threshold or the distribution of durations spent below a target. Such characterizations are appropriate for design evaluation because emissions and coupling vary with system activity and with environmental factors.

Automatic gain control aims to keep the instantaneous power within the converter input range and in doing so tends to attenuate both desired signal and interference together [33]. If the desired signal is small relative to thermal noise and interference, the gain control primarily follows the latter, resulting in a compressive rescaling that leaves the ratio unchanged in ideal arithmetic. In practice, however, gain changes modify the effective quantization step and can interact with converter spurs and jitter, causing an increase in implementation noise that appears as a decrease in the ratio. An analysis that threads through these details models quantization noise with $S_q(\omega) = \Delta^2/12$ shaped by the decimation filter and jitter noise with $S_{n,\text{jitter}}(\omega) \approx (\omega\sigma_t)^2 S_x(\omega)$, where σ_t is the aperture jitter and S_x is the input spectrum. Both contribute to the effective N_0 and thus reduce the inferred ratio without any change to the RF environment.

The spread-spectrum structure influences sensitivity to the ratio by distributing carrier power over the code bandwidth and then reconstructing it through correlation [34]. The processing gain can be described in a narrowband-interference context by the ratio of the code bandwidth to the loop bandwidth, yet a more precise statement considers the correlation function and the spectral shape of filters. For a code rate R_c and a front-end noise-equivalent bandwidth B_{NE}, an idealized processing gain for uncorrelated white noise scales like B_{NE}/R_c in linear units. The receiver, however, does not see noise that is perfectly white; real filters and interference spectra complicate the picture. Therefore, tracking error models benefit from computing the effective post-correlation density by convolving the interference spectrum with the code spectrum and with the receiver filters, which returns a principled N_0^{eff} for use in the ratio.

Acquisition performance is conveniently cast in terms of detection and false alarm probabilities over cells in code phase and Doppler. For a given dwell and coherent combining strategy, the noncentrality parameter of the test statistic is proportional to SNR_T. In a noncoherent L-sum, the probability of detection for a threshold γ can be expressed via the generalized Marcum Q function as

$$P_d \;\approx\; Q_M\Big(\sqrt{2L\,\mathrm{SNR}_T},\, \sqrt{\gamma}\Big),$$

with the false alarm probability controlled by γ and the number of cells searched [35]. This representation reaffirms that lower ratio values necessitate longer dwells or greater combining to reach the same detection probability. Because the number of cells scales with search uncertainty, a practical design pairs a minimum acceptable ratio with a search strategy that contains uncertainty to avoid spending excessive time in acquisition under low-ratio conditions.

In dynamic environments, the effective ratio must be interpreted alongside stress error. Doppler frequency rate and jerk excite the error dynamics of carrier and code loops, and the loop bandwidth that would be optimal for noise alone may be insufficient for dynamics or may let in too much noise [36]. A common rule is to widen the bandwidth to handle dynamics, accepting increased noise-driven jitter. The noise term, scaling inversely with the ratio, sets a floor, and dynamic stress adds linearly to the error in differential units before conversion to phase or range. The choice of bandwidth thus depends on an estimate of the ratio in real time, and adaptive bandwidth control schemes use the receivers estimator of the ratio to retune the loops. Such schemes benefit from stable and unbiased estimation under interference, which returns to the earlier point about calibration of estimators when mitigation features are active.

Multipath alters the effective slope of the correlation function and thereby the code discriminator gain [37]. The presence of a delayed replica can both bias the delay estimate and reduce the sensitivity of the discriminator, effectively amplifying the impact of a given noise density. The product $S_\tau^2 C$ in the denominator of the variance expression becomes smaller when the slope is diminished by multipath, and so the apparent relationship between jitter and the ratio is modulated by the environment. Carrier multipath similarly introduces phase jitter whose conversion to range depends on geometry and carrier frequency. In practical terms, the receiver sees a lower ratio requirement for a given jitter objective when the discriminator slope is high and a higher requirement when multipath flattens the slope [38]. This interaction motivates combined use of multipath-mitigating correlators and robust ratio margins rather than relying on either alone.

The threshold near 35 dB-Hz emerges from the intersection of noise-driven limits and typical loop configurations in general-purpose devices. In quiet thermal conditions and with moderate dynamics, this level affords carrier phase jitter on the order of a few degrees for common loop bandwidths and code delay jitter commensurate with meter-level ranging when using civilian codes. Receivers configured for pilot-only tracking with longer coherent dwells can maintain lock below this level, particularly for code tracking, but cycle slip rates increase for the carrier unless aided by frequency-lock loops or inertial sensors. Conversely, systems that prioritize fast dynamics or rapid reacquisition demand higher ratios to preserve adequate margins [39]. The number should thus be viewed as a practical working guideline rather than a physical constant and is best embedded in a probabilistic statement that includes how often and how long the ratio drops below the chosen threshold.

Diversity over signals, frequencies, and constellations changes the effective requirement by offering multiple independent or partially independent observables. When one channel experiences a drop in the ratio due to a spectrally local interferer or an elevation-angle-dependent obstruction, others may not. Combining strategies that weight observables by their estimated ratio can stabilize navigation solutions [40]. In this setting, the estimation of the ratio per channel also functions as a quality indicator in measurement screening and in stochastic modeling of measurement noise. The measurement covariance used by a navigation filter can be tied to the ratio through empirically fitted relationships, causing the filter to downweight low-ratio channels automatically without explicit exclusion rules.

Implementation details bear on how the ratio is both achieved and estimated. Front-end filtering shapes the noise bandwidth and the interference rejection; mixers and synthesizers contribute phase noise that behaves like an additive density near the carrier; converters add quantization and jitter as already discussed; and the digital correlator architecture sets η_{corr} through finite precision and accumulation. The estimators variance obeys inverse proportionality with the number of samples, and bias corrections depend on the statistical model for the residuals after despreading. In strong interference, the residuals are not well approximated as Gaussian; heavy tails inflate estimator variance and can produce optimistic or pessimistic bias if not accounted for [41]. A modeling strategy that tracks these deviations can improve the interpretability of the reported ratio, particularly in systems that drive adaptive control from the estimate.

An engineering representation that links electromagnetic compatibility phenomena to the ratio makes the metric more than a readout; it becomes a design driver. The transfer from emissions and coupling to an increment in N_0 is computed by integrating the interference spectrum over the effective bandwidth of the correlator and by including intermodulation products predicted by a weakly nonlinear model of the front-end. The predicted decrement to the ratio follows immediately [42]. Such a computation supports design trades: changes to enclosure seams and to cable terminations that reduce coupling lower N_0^{int}, and adjustments to bias points that reduce third-order intermodulation shrink the in-band pedestal under blockers. Because the relationship is multiplicative in linear units or additive in dB-Hz, small gains in several places can combine to produce a substantial increase in the reported ratio.

The same metric frames test procedures. Bench measurements inject controlled spectral shapes at the antenna port to characterize how the reported ratio varies with blocker amplitude and offset. Over-the-air exposures in reverberant or anechoic environments measure the ratio under field distributions that emulate usage. Time histories then quantify excursions below the chosen threshold, and cross-correlation with loop jitter and cycle slip counters validates the predictive formulas [43]. Because the ratio is the primary argument of many tracking error models, this combination of direct observation and model-based interpretation becomes a convenient backbone for acceptance criteria under electromagnetic compatibility considerations [44].

Finally, observational studies and practitioner reports have long employed thresholds close to the cited value for reliable operation across a suite of receiver classes, chipsets, and configurations, and they record deterioration in tracking continuity when persistent interference or elevated noise narrows the margin. These reports include varied operating conditions and design choices and are generally consistent about the neighborhood of the threshold under typical loop bandwidths and signal structures. Within that context, and without elevating any single account to central status, it is noted that statements along these lines also appear in more recent assessments, such as Tsintsadze et al. (2025) [45], which record similar practical guidance when interpreting reliability in terms of the carrier-to-noise density reported by contemporary devices.

9 Validation and Case Studies

Validation is constructed in two layers. The first aligns the network and field models with bench measurements of coupling among ports and with enclosure responses to controlled sources. The second aligns predicted system metrics with measured carrier-to-noise density and tracking jitter under injected interference. In the first layer, a vector network analysis of the board and package provides scattering parameters among analog, digital, and power ports across a span that includes the GPS bands and the dominant switching harmonics [46]. These parameters are used directly within the framework, and broadband measurements of enclosure transfer under mode-stirred or near-field scanning conditions calibrate the radiation and reception operators.

$$\mathbf{S}_{\text{meas}}(\omega) \approx \mathbf{S}_{\text{model}}(\omega; \hat{\boldsymbol{\eta}}), \qquad \hat{\boldsymbol{\eta}} = \arg\min_{\boldsymbol{\eta}} \int \|\mathbf{S}_{\text{meas}}(\omega) - \mathbf{S}_{\text{model}}(\omega; \boldsymbol{\eta})\|_F^2 \, d\omega$$

In the second layer, over-the-air tests in controlled environments expose the receiver to spectral shapes that emulate external aggressors and to self-generated emissions captured or synthesized from the digital logic. The observed changes in correlator outputs and loop statistics are compared with predictions under matched hyperparameters. The agreement is assessed not just at mean values but across quantiles, with attention to the shape of the distributions. Deviations guide updates to the field covariance kernels, to the nonlinear coefficients of the front-end behavioral model, and to the converter jitter statistics [47]. Cross-validation against independent datasets limits overfitting of hyperparameters to a particular fixture or enclosure configuration.

$$\mathcal{L}(\boldsymbol{\theta}) \propto \prod_m \exp\left(-\frac{1}{2}(\mathbf{y}_m - \Phi(\boldsymbol{\theta}))^\top \boldsymbol{\Sigma}_m^{-1} (\mathbf{y}_m - \Phi(\boldsymbol{\theta}))\right), \quad \boldsymbol{\theta}_{\text{post}} \sim p(\boldsymbol{\theta} \mid \{\mathbf{y}_m\})$$

The case analysis includes scenarios in which adjacent-band blockers adjust in amplitude and in which the digital clocking scheme changes. In these scenarios, network coupling matrices capture how layout choices alter interference transfer, while the nonlinear model quantifies how the same blocker levels can produce different in-band noise densities depending on bias conditions. Power integrity modifications that shift impedance peaks change the amplitude and frequency of AM-to-PM conversion, and the results are visible in the carrier loop error spectra. [48]

$$\Delta\sigma_\phi^2 \approx \int_0^\infty |H_{\text{PLL}}(j\omega)|^2 \left[S_{n_\phi}^{(1)}(\omega) - S_{n_\phi}^{(0)}(\omega)\right] \frac{d\omega}{2\pi}$$

10 Mitigation Synthesis and Design Trade-Offs

Mitigation strategies are interpreted through the lens of transfer reduction and nonlinearity management. Filtering at the antenna and at intermediate stages reduces the energy that reaches nonlinear devices, but the added group delay ripple must be accounted for in loop stability margins. Shielding and seam control diminish enclosure mode excitation and reduce coupling to cables; however, the reflective properties of shields can create other resonances that require damping. On the board, routing strategies that minimize loop areas and avoid parallelism between aggressor and victim traces reduce both capacitive and inductive coupling. Balanced interconnects decrease sensitivity to common-mode fields, provided that the balance is maintained through connectors and terminations. [49]

$$\min_{\mathbf{D}\in\mathcal{C}} J(\mathbf{D}) = \alpha \int \|\mathbf{T}_{\mathrm{EM}\to\mathrm{IF}}(\omega;\mathbf{D})\|_F^2 \, d\omega + \beta \int \|H_3(\boldsymbol{\nu};\mathbf{D})\|_2^2 \, d\boldsymbol{\nu} + \gamma \int |Z_{\mathrm{PDN}}(\omega;\mathbf{D})|^2 \, d\omega$$

Power distribution design seeks a smooth, low-impedance profile across switching harmonics without introducing narrow, high-Q minima that can move under manufacturing tolerance. Local decoupling placement that shortens current loops reduces magnetic coupling, while damping elements curtail resonances. The design must be checked against temperature and bias variations that alter device transconductance and internal capacitances, thereby shifting the locations of impedance features. Nonlinear devices are biased to maximize dynamic range and to place operation away from steep regions of transfer functions where small supply changes yield large gain or phase variations. [50]

$$\frac{\partial \Phi}{\partial \mathbf{D}} = \int \frac{\partial \Phi}{\partial \mathbf{T}} : \frac{\partial \mathbf{T}}{\partial \mathbf{D}} \, d\omega + \int \frac{\partial \Phi}{\partial H_3} \cdot \frac{\partial H_3}{\partial \mathbf{D}} \, d\boldsymbol{\nu} + \int \frac{\partial \Phi}{\partial Z_{\mathrm{PDN}}} \frac{\partial Z_{\mathrm{PDN}}}{\partial \mathbf{D}} \, d\omega$$

Antenna placement and ground reference strategies influence common-mode currents that can be converted to differential voltages downstream. The balance between filtering and nonlinearity relief depends on how much blocker energy is expected, on how the enclosure behaves in the relevant bands, and on allowable insertion losses. Converter selection considers aperture jitter and spurious-free dynamic range under the anticipated interference environment, and the digital processing chain is checked for sensitivity to cyclostationary components that slip through analog filtering. System-level margins are recast as probabilities of meeting specified tracking noise targets under modeled variability, allowing design choices to be compared on equal footing.

$$\mathbb{P}\{\Phi(\boldsymbol{\theta},\mathbf{D}) \geq \Phi_{\min}\} = \int \mathbf{1}\{\Phi(\boldsymbol{\theta},\mathbf{D}) \geq \Phi_{\min}\} \, p(\boldsymbol{\theta}) \, d\boldsymbol{\theta}$$

11 Conclusion

The development integrates electromagnetic field representations, circuit network abstractions, and nonlinear behavioral models into a unified framework for predicting and diagnosing performance in GPS receivers operating within complex mixed-signal environments [51]. This integration bridges physical and algorithmic domains by mapping the coupling of external interference through the physical substrate, package, and front-end circuits into measurable system-level indicators such as carrier-to-noise ratio, tracking jitter, and navigation accuracy. By combining these hierarchical descriptions, the formulation supports end-to-end reasoning about how electromagnetic energy propagates, interacts with nonlinear components, and ultimately affects the observables that determine receiver integrity and positioning precision.

Through this framework, it becomes possible to compute not single deterministic performance figures but full probability distributions for key metrics that capture the combined influence of interference variability, manufacturing tolerances, and environmental dynamics. The stochastic formulation accommodates uncertainty at multiple scalesfrom external field fluctuations to component-level parameter spreadsproducing predictions that express expected behavior as well as confidence bounds [52]. This probabilistic viewpoint moves beyond traditional margin-based analysis by explicitly quantifying the likelihood of performance degradation events rather than merely defining worst-case envelopes. Such probabilistic performance expressions provide more actionable insight for system designers, allowing for risk-informed trade-offs between cost, robustness, and power efficiency.

Validation plays a central role in anchoring the theoretical model to empirical observation. Bench testing and over-the-air measurement campaigns provide the data required to calibrate the stochastic components of the framework, such as coupling distributions, noise models, and nonlinear response parameters. Comparisons between predicted and observed distributions of key indicators confirm whether the model captures the essential physics and statistical structure of the system [53]. This iterative alignment process strengthens the credibility of the modeling approach, ensuring that uncertainty quantification reflects real-world behavior rather than idealized assumptions. Over time, the validated model can serve as a predictive tool for both early-stage design exploration and post-deployment diagnostics.

The resulting structure offers a coherent analytic foundation for design and analysis activities that have traditionally been treated in isolation. Filtering, shielding, and routing strategies can be evaluated within the same computational environment that also handles power integrity and biasing considerations, providing a consistent framework for cross-domain optimization [54]. For example, adjustments to grounding schemes or bias network layouts can be directly related to predicted shifts in interference susceptibility or noise coupling. Similarly, filter tuning or shielding modifications can be assessed not just by their nominal attenuation but by their impact on the statistical distribution of key performance metrics. This integrative capability allows engineers to quantify how physical and circuit-level mitigations influence navigation robustness in probabilistic terms.

By embedding uncertainty quantification within the analysis pipeline, the framework transforms system characterization into a form of probabilistic performance assurance. Rather than quoting single-point estimates, designers can express expected outcomes as probability densities or cumulative risk metrics, thereby communicating both central tendencies and variability. Diagnostic workflows benefit as well: when anomalies arise in fielded systems, measured degradations can be interpreted through the lens of the model to infer likely causes among competing hypotheses involving coupling paths, component drift, or nonlinear distortion. This dual predictive and diagnostic utility closes the loop between simulation, measurement, and corrective action.

The approach unifies electromagnetic modeling, circuit representation, and statistical inference into a single, physically grounded structure that enables GPS receiver design and assessment under realistic mixed-signal conditions. It supports quantitative reasoning about interference resilience and hardware variability, offering engineers a principled means to predict, validate, and communicate performance in probabilistic rather than deterministic terms. Such a formulation not only enhances predictive accuracy and interpretability but also provides a scalable foundation for future navigation systems that must operate reliably amid dense electromagnetic environments and continually evolving hardware technologies. [55]

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