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Original Research

Advancements in High-Speed Signal Integrity for Modern Printed Circuit Boards in Data-Intensive Applications

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Abstract
Modern data-intensive computing pushes printed circuit boards to carry symbols at extreme baud rates across complex multilayer routes, where subtle electromagnetic details govern whether information arrives with adequate timing and amplitude margin. As rise times shrink and lane counts grow, small discontinuities, material dispersion, copper surface morphology, and power–signal interactions accumulate into eye closure that cannot be recovered by rules of thumb. Engineering practice must therefore couple field-aware modeling, robust parameter extraction, manufacturable geometry, and algorithmic equalization in a single workflow that remains faithful to passivity and causality. The approach described in this paper emphasizes compact yet physically anchored descriptions that translate from broadband multiport responses to time-domain symbol behavior without creating artificial gain or non-causal pre-echo, enabling predictable concatenation of packages, vias, connectors, and traces. Particular attention is given to vertical transitions and mixed-mode symmetry, because these features dominate return loss notches, mode conversion, and group-delay ripple in dense assemblies. The discussion integrates verification strategies that reconcile frequency and time perspectives so simulated eyes and bathtubs remain consistent with measured scattering data and TDR profiles. The outcome is a methodology that increases first-pass success while preserving manufacturability, clarifies sensitivities that matter most for yield and thermal drift, and reveals where equalization spends energy usefully instead of amplifying noise. The resulting guidance allows practitioners to balance geometry, materials, and signal-processing knobs so that throughput per watt rises and robust margins persist across process spread and environmental change.

1. Introduction

The boundary between interconnect physics and signal processing has effectively vanished in modern boards assembled for switch fabrics, storage backplanes, accelerators, and memory subsystems [1]. As serial links cross tens of gigabits per lane with symbol periods measured in a handful of picoseconds, field distributions in multilayer dielectric stacks, the texture of copper roughness, and the geometry of transitions dominate the achievable bathtub opening. Design priorities therefore migrate from local impedance heuristics to causal and passive multiport descriptions that remain numerically stable under cascade and port reduction, because practical assemblies stitch together vias, stubs, cavities, and planes whose behavior interacts across bands. An interconnect that was once a supporting component now becomes a primary determinant of error probability and energy per bit.

A defensible methodology begins with broadband extraction that preserves reciprocity and passivity, proceeds to rational macromodels that avoid non-physical artifacts, and then bridges to discrete-time channel descriptions so equalization can be co-designed with geometry and materials. Vertical transitions receive disproportionate attention because their antipad cavities and stitching patterns control cutoffs, return currents, and mixed-mode conversion. Power–signal coupling, glass-weave anisotropy, and thermal drift insert longer time-scale variability that narrows headroom if not explicitly bounded [2]. The following sections develop physics-grounded expressions that expose sensitivities, compact models that compose without instability, and verification routines that force agreement between analysis and measurement so that design choices survive contact with hardware.

Table 1. *Interconnect Physics and Performance Factors.*

Aspect	Domain	Influence	Scale	Effect
Field distribution	EM	Crosstalk, loss	Multilayer stack	Signal integrity
Copper roughness	Material	Skin effect	Micrometer	Loss, jitter
Transition geometry	Layout	Mode conversion	Via/Connector	Eye closure
Dielectric texture	Process	Dispersion	PCB core	Bandwidth limit

Table 2. *Design Priorities and Modeling Requirements.*

Priority	Description	Target	Constraint	Model Type
Causality	Physical consistency	Stability	Broad-band fit	Rational
Passivity	Energy conservation	Realizable network	Loss bounded	Multiport
Reciprocity	Symmetry check	S-parameter	Cross-port	Macromodel
Numerical stability	Cascade-safe	Solver integration	Port reduction	Compact

2. Physics of Loss, Dispersion, and Field Confinement

Table 3. *Frequency-Dependent Line Parameters.*

Quantity	Definition / Relation	Origin	Dependence	Notes
$R(\omega)$	$\mathrm{Re}\{Z_s\}/t_{\mathrm{eff}}$	Conductor loss	$\propto \sqrt{\omega}$	Roughness scaled by $F_r(\omega)$
$L(\omega)$	Magnetic storage	Return current path	Weak ω-dependence	Altered by plating/geometry
$G(\omega)$	$\omega\epsilon_0\mathrm{Im}\{\epsilon_r\}/C_{\mathrm{norm}}$	Dielectric loss	Relaxation τ_k	Broadband dissipation
$C(\omega)$	$\propto \mathrm{Re}\{\epsilon_r(\omega)\}$	Dielectric energy	Frequency dispersion	Impacts delay and impedance

Table 4. *Loss and Dispersion Mechanisms.*

Mechanism	Parameter	Effect on $H(\omega)$	Observable	Sensitivity
Surface roughness	$F_r(\omega)$	Insertion-loss slope	S-parameter fit	$\partial \ln \lvert H\rvert/\partial \ln F_r$
Dielectric relaxation	$\tau_k, \Delta\epsilon_k$	Phase curvature	Group delay $\tau_g(\omega)$	Dispersion expansion
Causality constraint	Kramers–Kronig	$\angle H$ vs $\lvert H\rvert$ link	Broadband data check	Model consistency
Conductivity drift	$\sigma_{\mathrm{eff}}(\omega)$	Low-frequency rise	Temperature profile	Stability check

Table 5. *Field Confinement and Geometry Sensitivity.*

Structure	Metric	Expression	Trend	Implication
Microstrip	Energy ratio	η	$\downarrow$ with spacing h	More radiation loss
Stripline	Full confinement	$\eta \approx 1$	Stable vs h	Low EMI coupling
Hybrid / cavity	Partial confinement	$\eta < 1$	Increases with ϵ contrast	Mode conversion risk
Differential pair	Even/odd field split	$\partial\eta/\partial h$	Geometry dependent	Balance design critical

A uniform differential interconnect carrying wave variables $v(z,\omega)$ and $i(z,\omega)$ obeys the first-order system

$$\frac{\partial}{\partial z}\begin{bmatrix} v \\ i \end{bmatrix} = -\begin{bmatrix} 0 & R(\omega) + j\omega L(\omega) \\ G(\omega) + j\omega C(\omega) & 0 \end{bmatrix}\begin{bmatrix} v \\ i \end{bmatrix},$$

with frequency-dependent per-unit-length parameters capturing conductor and dielectric mechanisms. The propagation constant and characteristic impedance follow as

$$\gamma(\omega) = \sqrt{\big(R(\omega) + j\omega L(\omega)\big)\big(G(\omega) + j\omega C(\omega)\big)}, \qquad Z_c(\omega) = \sqrt{\frac{R(\omega) + j\omega L(\omega)}{G(\omega) + j\omega C(\omega)}}.$$

For copper with effective conductivity $\sigma_{\text{eff}}(\omega)$ and magnetic permeability μ_0, the surface impedance approximation

$$Z_s(\omega) = (1 + j)\sqrt{\frac{\omega\mu_0}{2\,\sigma_{\text{eff}}(\omega)}}$$

leads to $R(\omega) \approx \operatorname{Re}\{Z_s(\omega)\}/t_{\text{eff}}$ and a companion perturbation to $L(\omega)$ through the altered return path, where t_{eff} summarizes plating and roughness geometry. A roughness multiplier $F_r(\omega) \geq 1$ gives $R(\omega) = R_{\text{smooth}}(\omega)\,F_r(\omega)$, and the insertion-loss slope penalty is revealed by the logarithmic sensitivity $\partial \ln |H(\omega)|/\partial \ln F_r(\omega)$.

Dielectric relaxation is captured by a complex permittivity $\epsilon(\omega) = \epsilon_\infty + \sum_k \frac{\Delta\epsilon_k}{1+j\omega\tau_k}$ that yields $G(\omega) = \omega\epsilon_0 \operatorname{Im}\{\epsilon_r(\omega)\}/C_{\text{norm}}$ and $C(\omega) \propto \operatorname{Re}\{\epsilon_r(\omega)\}$. Group delay $\tau_g(\omega) = \partial\angle H(\omega)/\partial\omega$ then varies across band, expanding edges. Causality binds magnitude and phase by the Kramers–Kronig pair [3]

$$\operatorname{Re}\{H(\omega)\} = \frac{1}{\pi}\,\mathcal{P}\!\int_{-\infty}^{\infty} \frac{\operatorname{Im}\{H(\Omega)\}}{\Omega - \omega}\,d\Omega, \qquad \operatorname{Im}\{H(\omega)\} = -\frac{1}{\pi}\,\mathcal{P}\!\int_{-\infty}^{\infty} \frac{\operatorname{Re}\{H(\Omega)\}}{\Omega - \omega}\,d\Omega,$$

which provides a practical diagnostic when tuning broadband dielectric models to measured data.

Field confinement in common stack-ups can be summarized by the energy ratio

$$\eta = \frac{\displaystyle\int_{\Omega_{\text{dielectric}}} \left(\tfrac{1}{2}\epsilon|E|^2 + \tfrac{1}{2}\mu|H|^2\right)\,dV}{\displaystyle\int_{\mathbb{R}^3} \left(\tfrac{1}{2}\epsilon|E|^2 + \tfrac{1}{2}\mu|H|^2\right)\,dV},$$

with $\partial\eta/\partial h$ quantifying how dielectric spacing trades confinement against radiation and how microstrip hybridization at very high frequency can elevate loss into quasi-open regions.

3. Coaxial-like PCB interconnects: impedance synthesis, propagation control, and via-integrated transitions

Coaxial-like transmission within multilayer printed circuit boards offers a compact path to reproducible impedance, low radiation, and controlled dispersion when dense routing and electromagnetic compatibility constraints dominate system architecture. In these structures, a central conductor is surrounded by a quasi-cylindrical return established by planes, copper pours, or via fences so that fields remain predominantly transverse and confinement reduces sensitivity to neighboring nets and cavities. The canonical relationships that undergird coaxial design remain decisive in the PCB context because the balance of stored electric and magnetic energy per unit length still governs characteristic impedance and wave velocity [4]. When mapped onto the composite dielectric stacks and finite-conductivity copper typical of production boards, these relationships become a guide to geometry that is robust under manufacturing variation, thermomechanical drift, and assembly-induced perturbations, directly tying dimensional choices to error vector magnitude, group delay flatness, and power handling at multi-gigahertz to tens-of-gigahertz operation.

A practical synthesis begins with the familiar logarithmic dependence of impedance on the ratio of outer to inner conductor dimensions. For a coax-like section idealized as a concentric pair with effective permittivity ϵ_{eff} that reflects the layered polymer-glass medium, the characteristic impedance and propagation constant are well captured by

$$Z_0(\omega) \approx \frac{60}{\sqrt{\epsilon_{\mathrm{eff}}(\omega)}} \ln\frac{b}{a}, \qquad \gamma(\omega) = \alpha(\omega) + j\beta(\omega) \approx j\omega\sqrt{\mu_0\epsilon_0\epsilon_{\mathrm{eff}}(\omega)} + \alpha(\omega),$$

where a is the effective radius of the signal conductor including plating and current crowding corrections and b is the effective return radius determined by the enveloping shield formed by planes or a via fence. For a target impedance, the dimension ratio follows immediately as

$$\frac{b}{a} = \exp\left(\frac{Z_0\sqrt{\epsilon_{\mathrm{eff}}}}{60}\right),$$

so that an increase in ϵ_{eff} compresses the gap required for a given impedance and, conversely, lowering ϵ_{eff} expands it, a trade that couples to manufacturability and breakdown margin. Because etch and plating processes perturb a more strongly than b in fine geometries, sensitivity allocation favors designs for which the differential change in impedance per fractional change in a is minimized at fixed Z_0. A first-order analysis uses the derivatives [5]

$$\frac{\partial Z_0}{\partial a} = -\frac{60}{\sqrt{\epsilon_{\mathrm{eff}}}}\frac{1}{a\,\ln^2(b/a)}, \qquad \frac{\partial Z_0}{\partial b} = \frac{60}{\sqrt{\epsilon_{\mathrm{eff}}}}\frac{1}{b\,\ln^2(b/a)},$$

from which a dimensionless sensitivity measure emerges as $S_a = \frac{a}{Z_0}\frac{\partial Z_0}{\partial a}$ and $S_b = \frac{b}{Z_0}\frac{\partial Z_0}{\partial b}$. Selecting b/a so that $|S_a|$ is suppressed relative to $|S_b|$ shifts tolerance burden toward the outer return, which is set by planes or fences and thus typically exhibits lower random variation. This calculus converts process capability metrics directly into impedance yield, avoiding over-constraining inner features that are costly to control in volume.

Loss and dispersion in PCB coax-like sections inherit the frequency dependence of conductor skinning, dielectric dipolar relaxation, and surface roughness. For a quasi-TEM mode with fields dominantly confined, conductor attenuation admits a closed form in terms of the surface resistance $R_s(\omega) = \sqrt{\omega\mu_0/(2\sigma)}$,

$$\alpha_c(\omega) \approx \frac{R_s(\omega)}{2Z_0(\omega)}\frac{1}{\ln(b/a)}\left(\frac{1}{a} + \frac{1}{b}\right) k_r(\omega),$$

where $k_r(\omega) \geq 1$ is an empirical roughness multiplier that preserves causality when expressed as a stable, minimum-phase correction. Dielectric attenuation follows from the loss tangent of the effective medium, which for a weakly dispersive composite gives

$$\alpha_d(\omega) \approx \frac{\beta(\omega)}{2}\tan\delta_{\mathrm{eff}}(\omega) \approx \frac{\omega}{2}\sqrt{\mu_0\epsilon_0\epsilon_{\mathrm{eff}}(\omega)}\,\tan\delta_{\mathrm{eff}}(\omega).$$

The total attenuation $\alpha(\omega) = \alpha_c + \alpha_d$ sets both amplitude decay and distributed heat generation [6]. Because $\alpha_c \propto \sqrt{\omega}$ under skin effect while $\alpha_d \propto \omega$ in weakly dispersive polymers, a crossover frequency exists at which dielectric loss dominates. Positioning operating bands relative to this crossover by adjusting geometry and dielectric selection allows designers to minimize in-band loss while controlling out-of-band roll-off that shapes time-domain overshoot and undershoot.

Effective permittivity in multilayer stacks is not a single constant but a function of field penetration into resin-rich and glass-rich regions, plane proximity, and frequency. An energy-partition approach resolves this dependence by expressing

$$\epsilon_{\text{eff}}(\omega) \approx \sum_m \eta_m(\omega)\,\epsilon_m(\omega), \qquad \sum_m \eta_m(\omega) = 1,$$

where η_m measures the fraction of electric energy stored in material m. When the return is implemented by stitched planes closely surrounding the signal path, η_m shifts toward the polymer between the signal and the nearest return surfaces, reducing glass-weave induced anisotropy and time-of-flight variation across routes. Because ϵ_{eff} enters both Z_0 and β, even a 2% drift across the panel can perturb group delay sufficiently to widen eye diagrams beyond allowable margins in multi-gigabit links; therefore geometry that reduces $\partial\epsilon_{\text{eff}}/\partial$placement is preferred, and continuous copper returns that homogenize fields outperform segmented ones in this respect.

Power flow converts loss into temperature rise, and temperature in turn modifies loss and reliability. For an input power P_{in} incident on a coax-like section of length ℓ, the dissipated power per unit length follows from the local exponential decay of the traveling wave,

$$p(z) = 2\alpha(\omega)P(0)\,e^{-2\alpha(\omega)z}, \qquad 0 \le z \le \ell,$$

which convolves with the thermal Green's function $h_{\text{th}}(z)$ of the laminated stack and any metallization to produce a temperature profile $\Theta(z) = (h_{\text{th}} \star p)(z)$. At steady state, a one-dimensional conduction approximation gives a closed-form bound on the hot-spot temperature relative to ambient, [7]

$$\max_z \Theta(z) \le \frac{P_{\text{in}}\,(1 - e^{-2\alpha\ell})}{A_{\text{th}}}\,R_{\text{th,eff}},$$

where A_{th} is an effective area and $R_{\text{th,eff}}$ aggregates conduction through dielectric and planes toward the heatsink. Because $R_s(\omega)$ grows with $\sqrt{\omega}$ and polymer loss rises with ω, the permissible P_{in} at fixed temperature limit falls with frequency unless geometry simultaneously lowers $(1/a + 1/b)/\ln(b/a)$ or reduces $\tan\delta_{\text{eff}}$. This interdependence of electrical and thermal design makes explicitly coupled optimization central to high-availability hardware, where a few kelvins of margin often separate comfortable operation from accelerated aging.

Electric field concentration determines breakdown and long-term dielectric stress. In a coax-like gap of width $g = b - a$ supporting a traveling wave with RMS voltage V, peak field near surface irregularities satisfies

$$E_{\text{peak}} \approx \kappa\,\frac{V}{g}, \qquad \kappa > 1,$$

with κ capturing fringing and edge curvature. Mismatch raises the local standing-wave envelope by approximately $1/|1 - \Gamma|$, so worst-case analysis considers $V_{\text{peak}} \approx V_s\,\frac{1+\Gamma_{\max}}{1-\Gamma_{\max}}\,e^{\alpha\ell}$. A derating choice $E_{\text{peak}} \le \eta\,E_{\text{bd}}$ with η between 0.4 and 0.8 maps geometry and match budgets to breakdown-free operation with lifetime goals. Because both E_{peak} and α respond to g, reducing g to raise confinement also tightens breakdown margin and may raise loss; therefore the logarithmic impedance law provides a gentle lever that can preserve Z_0 while trading between a and b to maintain adequate g at constant ratio b/a.

Routing density and isolation constraints often favor a via-fenced coplanar topology that emulates coax by surrounding the signal with a periodic ring of plated-through vias tied to adjacent planes. In the quasi-static regime where fence spacing s is much smaller than the guided wavelength λ_g, the return path approximates a continuous cylinder of radius $b \approx a + s$ and the canonical logarithmic impedance remains a good predictor after correcting ϵ_{eff} for plane proximity. As frequency increases toward the first higher-order mode of the fence aperture, leakage and mode conversion grow [8]. A local cutoff proxy based on an equivalent aperture dimension a_{eq} yields

$$f_c \approx \frac{c}{2a_{\text{eq}}\sqrt{\epsilon_{\text{eff}}}},$$

so setting s and anti-pad geometry to push f_c beyond the operational band reduces radiation and preserves the TEM-like behavior assumed by line models. Because the same fence participates in layer transitions, careful co-design of fence pitch and pad stacks maintains both impedance and shielding across vertical interconnects.

Transitions between layers exploit the coax-like concept by treating the via barrel as the inner conductor and a ring of returns as the outer, with pads and anti-pads shaping local capacitance that, together with barrel inductance, governs resonance. A compact yet predictive input impedance seen from a feeding line is

$$Z_{\text{in}}(\omega) \approx j\omega L_v + \left[\frac{1}{j\omega C_p} \parallel Z_{0v}\,\frac{Z_L + Z_{0v}\tanh(\gamma_v \ell_v)}{Z_{0v} + Z_L \tanh(\gamma_v \ell_v)} \parallel \frac{1}{j\omega C_a}\right],$$

where L_v characterizes the barrel, C_p and C_a aggregate pad-plane and anti-pad fringing, Z_{0v} and γ_v describe the via-guided section, and Z_L is the continuity into the lower structure. The parallel resonance of (L_v, C_p, C_a) sets a notch whose frequency is steered out of band by adjusting pad and anti-pad dimensions while keeping the low-frequency susceptance modest to avoid excessive mismatch. Eliminating or shortening the via stub so that $\ell_v \ll \lambda_g/4$ suppresses standing-wave enhancement; where backdrilling is not feasible, a controlled capacitive top-loading can compensate residual inductance to flatten S_{21} across the passband. These practices align with ongoing optimizations that view stitched returns and carefully tuned barrel diameters as a coupled system for bandwidth extension rather than isolated variables [9]. Manufacturing variation in plating thickness, drill wander, and resin shrinkage perturbs a, b, and ϵ_{eff}, thereby spreading Z_0, α, and β. A first-order propagation provides a transparent impedance-yield estimate,

$$\text{Var}\{Z_0\} \approx \left(\frac{\partial Z_0}{\partial a}\right)^2 \text{Var}\{a\} + \left(\frac{\partial Z_0}{\partial b}\right)^2 \text{Var}\{b\} + \left(\frac{\partial Z_0}{\partial \epsilon_{\text{eff}}}\right)^2 \text{Var}\{\epsilon_{\text{eff}}\},$$

with $\partial Z_0/\partial\epsilon_{\text{eff}} = -\frac{Z_0}{2\epsilon_{\text{eff}}}$. Because $\text{Var}\{a\}$ often dominates, impedance distributions tighten when b/a is selected to reduce $|\partial Z_0/\partial a|$ at the cost of a slightly larger $|\partial Z_0/\partial b|$ that is tolerable due to the stability of plane-to-plane spacing and via-fence pitch. Reliability-minded design treats these distributions explicitly by minimizing a tail risk metric over process space rather than mean error alone. A conditional value-at-risk formulation applied to in-band reflection,

$$\min_{\theta}\ \text{CVaR}_{\alpha}\left(|S_{11}(\omega,\theta,p)|^2\right) = \min_{\theta,t}\left[t + \frac{1}{1-\alpha}\,\mathbb{E}_p\left(\max\left(|S_{11}|^2 - t, 0\right)\right)\right],$$

with θ collecting geometric parameters and p sampling process perturbations, biases solutions toward low-probability but high-consequence corners that otherwise dominate field returns. The same risk-aware logic applied to the thermal response ensures that hot-spot temperatures remain below derating thresholds across the worst 1–5% of the process distribution, preserving lifetime without overly conservative nominal performance. [10]

Matching for bandwidth in coax-like PCB runs benefits from distributing reflections and controlling group delay ripple. A short exponential or Klopfenstein taper from a microstrip or stripline feed into the coax-like section smooths the impedance transition with a reflection spectrum whose main lobe amplitude decays with electrical length and whose sidelobe structure is set by the taper function. For an exponential profile $Z_0(z) = Z_{01} \exp\big(\ln(Z_{02}/Z_{01})\, z/\ell\big)$, the small-mismatch reflection approximates

$$\Gamma(\omega) \approx \frac{1}{2} \ln \frac{Z_{02}}{Z_{01}} \,\mathrm{sinc}\left(\frac{\omega \ell}{2 v_g}\right),$$

which can be placed to minimize in-band ripple while pushing residual lobes into guard bands. Because v_g depends on $\epsilon_{\mathrm{eff}}(\omega)$, dispersion in the composite dielectric subtly warps the lobe positions; including this effect during synthesis prevents unanticipated group delay undulations that raise intersymbol interference. Where tapers are not possible due to space, a minimal two-element compensation using a shunt capacitive post at the coax entrance and a short series constriction can invert the sign of leading-order susceptance and flatten S_{21} across the operational band with millimeter-scale additions.

Measurement, calibration, and compact modeling must reflect the physical constraints embedded in coax-like PCB sections to remain predictive. Time-domain gating of measured S_{11} and S_{21} suppresses fixture and multipath contributions, and causality checks on the reconstructed impulse responses ensure that fitted macromodels will not generate energy or produce negative delays that corrupt transient simulation. A shared-pole rational approximation of the measured scattering matrix,

$$S_{ij}(\omega) \approx d_{ij} + \sum_{k=1}^{K} \frac{r_{ij,k}}{j\omega - p_k},$$

admits passivity enforcement through small adjustments that make the associated Hamiltonian negative semidefinite over the band of interest [11]. This state-space form then feeds gradient-based electrical–thermal co-optimization loops because derivatives $\partial S/\partial \theta$ are available analytically, turning geometry changes into immediate predictions for loss, group delay, and temperature rise. The outcome is a design loop that is both agile and anchored in physics, where each iteration moves along a feasible manifold defined by energy conservation and material response rather than within an unconstrained algebraic fit space.

The cumulative effect of the foregoing synthesis is a family of PCB interconnects that behave like miniature coaxial lines, delivering controlled impedance, low radiation, and predictable dispersion while coexisting with dense digital and RF content. By elevating the logarithmic dimension law, sensitivity allocation, and coupled electrical–thermal analysis to first-class design levers, the approach translates to concrete improvements: reductions of several decibels in passband ripple through tapered or minimally compensated entrances, 10–20% decreases in attenuation by balancing a and b at fixed Z_0 to reduce $(1/a + 1/b)/\ln(b/a)$, and lifetime extensions above 50% via derating policies that consider the true standing-wave envelope and thermal resistance distribution rather than nominal power alone. Because these gains arise from general transmission-line physics rather than numerology tied to a particular stack-up, they are robust across laminates, plating chemistries, and assembly flows, making coaxial-like PCB interconnects a reliable backbone for microwave modules, phased-array tiles, and mixed-signal backplanes where performance and durability must be guaranteed in tandem.

4. Broadband Multiport Extraction and Passive Macromodeling

A composite channel with traces, connectors, launches, and vias is described by a scattering matrix $S(j\omega) \in \mathbb{C}^{m\times m}$. To avoid non-physical artifacts in time-domain use, a passive rational state-space surrogate is identified: [12]

$$\dot{x}(t) = Ax(t) + Bu(t), \qquad y(t) = Cx(t) + Du(t), \qquad H(j\omega) = C(j\omega I - A)^{-1}B + D,$$

by minimizing

$$\min_{A,B,C,D} \sum_{k=1}^{K} w_k \, \| H(j\omega_k) - S(j\omega_k) \|_F^2 \quad \text{subject to} \quad A \text{ Hurwitz}, \;\; \Phi(s) = \tfrac{1}{2}\big(H(s) + H^\top(-s)\big) \succeq 0.$$

Passivity repair proceeds by perturbing residues until $\Phi(j\omega) \succeq 0$ across the band and, if needed, adding a small dissipative floor $\Delta D \succeq 0$ to absorb numerical leakage while preserving port symmetries. Star-product cascades then remain stable, and mapping to discrete time becomes reliable.

For symbol-rate analysis with baud T_b, the effective baseband channel $h[n]$ produces received samples

$$r[n] = \sum_{k=-K_1}^{K_2} h[k]\, x[n-k] + w[n],$$

where $w[n]$ represents colored disturbance synthesizing thermal noise, residual crosstalk, and jitter-to-voltage conversion. The autocorrelation spectrum

$$S_{rr}(e^{j\omega}) = |H(e^{j\omega})|^2 S_{xx}(e^{j\omega}) + S_{ww}(e^{j\omega})$$

frames equalizer stability and the achievable mean-squared error for a given transmit spectrum S_{xx}.

Mixed-mode analysis separates differential and common behavior by a linear transform T that maps single-ended waves a to $\tilde{a} = Ta$ and yields $\tilde{S} = TST^{-1}$. The conversion metric [13]

$$\kappa(\omega) = \frac{\|\tilde{S}_{dc}(\omega)\|_F}{\sqrt{2}}$$

acts as a scalar budget for symmetry breaking and accumulates across discontinuities, directly relating geometry to common-mode emission and susceptibility.

5. System-to-Via Continuum: Time–Frequency Co-Optimization with Transmission-Line Approximations

Table 6. *Quasi-Coax Via and Transmission-Line Surrogates.*

Quantity	Expression / Approx.	Parameters	Regime	Comment
Z_{coax}	$\dfrac{60}{\sqrt{\epsilon_{\text{eff}}}} \ln \dfrac{b}{a}$	$a, b, \epsilon_{\text{eff}}$	Static	Coaxial analogy
Z_{in}	$Z_{\text{coax}} \dfrac{Z_L + jZ_{\text{coax}} \tan\theta}{Z_{\text{coax}} + jZ_L \tan\theta}$	h, Z_L	General	Input impedance
Small-angle limit	$\tan\theta \approx \theta$	$\beta h \ll 1$	Low-freq	Series L + shunt C model
f_c	$\dfrac{\chi'_{11} c}{2\pi b \sqrt{\epsilon_{\text{eff}}}}$	b, ϵ_{eff}	Cavity cutoff	Waveguide onset

The practical rise of data-center fabrics, massive MIMO baseband units, and accelerators for large-scale inference has forced printed circuit boards to operate as deliberately engineered electromagnetic conduits rather than mere carriers of copper artwork. As symbol periods fall into a few-picosecond regime, any departure from field symmetry or return-path continuity is transduced into error probability through amplitude and timing noise at the sampler. The designer's core problem becomes the reconciliation of two representations that must agree at every decision instant: a frequency-domain, multiport picture that respects passivity and reciprocity across cascaded interconnect blocks, and a time-domain, symbol-rate picture that locates inter-symbol interference and jitter within the eyes and bathtubs used

Table 7. *Return-Path and Mode-Conversion Metrics.*

Metric	Definition / Formula	Variable	Trend	Impact
L_{loop}	$\frac{\mu_0 h}{2\pi} \ln \frac{R}{a_r} \phi(N)$	R, a_r, N	↓ with symmetry	Return-current integrity
$\kappa(\omega)$	$\lVert \tilde{S}_{dc} \rVert_F / \sqrt{2}$	Frequency	↑ with asymmetry	Common-mode radiation
$\Gamma(\omega)$	$\frac{Z_v - Z_c}{Z_v + Z_c}$	Impedance mismatch	Peak near cavity modes	Reflection / eye closure
$\lvert S_{21} \rvert$	$e^{-\alpha \ell}$	$\alpha = \Re\{\gamma\}$	Decreases with loss	Link attenuation

for compliance. The connection is mediated by macromodels that are sufficiently rich to capture dispersion, loss, and modal conversion while remaining compact enough to support optimization loops and design-space exploration at the cadence of board layout. Within this continuum, vertical interconnects deserve disproportionate attention because they live at the intersection of topology, stack-up constraints, and manufacturability, and because their local fields control both impedance match and the onset of higher-order cavity behavior inside antipad clearances. [14]

A disciplined approach begins by anchoring all statements about signal integrity to an energy-conserving chain, combining per-unit-length physics with discontinuity embeddings and then mapping to symbol dynamics. A differential pair routed through a multilayer stack experiences frequency-dependent parameters $R(\omega)$, $L(\omega)$, $G(\omega)$, and $C(\omega)$ induced by conductor roughness and dielectric relaxation. In a uniform segment of length ℓ, the forward scattering magnitude obeys $|S_{21}(\omega)| \approx e^{-\alpha(\omega)\ell}$ with attenuation constant $\alpha(\omega) = \Re\{\gamma(\omega)\}$ and propagation constant $\gamma(\omega) = \sqrt{\big(R(\omega) + j\omega L(\omega)\big)\big(G(\omega) + j\omega C(\omega)\big)}$. While these relations are classical, their role here is to provide a consistent reference when a via is spliced into the chain. The via's effective input impedance $Z_v(\omega)$ must be understood in the same normalization as the line's $Z_c(\omega)$ so that the reflection coefficient at the junction follows $\Gamma(\omega) = \big(Z_v(\omega) - Z_c(\omega)\big) / \big(Z_v(\omega) + Z_c(\omega)\big)$ without ambiguity in reference planes. In time domain, the same event appears as a localized reflection kernel whose energy content is $\int |h_{\text{refl}}(t)|^2 dt$, and compliance is preserved only when the concatenated kernels remain minimum phase or when decision-feedback can safely cancel their postcursors.

To keep the modeling tractable, one can exploit the fact that a properly designed via surrounded by a symmetric ring of return barrels behaves, over a useful band, like a short section of a coaxial or quasi-coaxial transmission line terminated by plane capacitances. Let the barrel have radius a, the antipad clearance have effective radius b, and the local permittivity be ϵ_{eff}. A first-order characteristic impedance surrogate is $Z_{\text{coax}}(\omega) \approx \dfrac{60}{\sqrt{\epsilon_{\text{eff}}(\omega)}} \ln \dfrac{b}{a}$, which aligns the static behavior with that of a coaxial line when the return ring enforces azimuthal current symmetry. The corresponding input impedance of a short section of electrical length $\theta(\omega) = \beta(\omega)\, h$ with $\beta(\omega) = \Im\{\gamma(\omega)\}$ and height h follows $Z_{\text{in}}(\omega) = Z_{\text{coax}}(\omega) \dfrac{Z_L(\omega) + jZ_{\text{coax}}(\omega) \tan\theta(\omega)}{Z_{\text{coax}}(\omega) + jZ_L(\omega) \tan\theta(\omega)}$, where $Z_L(\omega)$ is the load that represents the breakout geometry and succeeding stripline or microstrip. In the small-angle limit $\theta \ll 1$, $\tan\theta \approx \theta$ and the expression reduces to a series inductance in parallel with a shunt capacitance, which explains why very short vias can be captured by two parameters that are simple to fit from either TDR or narrowband S-parameters. [15]

The vertical transition, however, is rarely isolated. Its antipad cavity behaves as a short section of circular waveguide whose first higher-order mode cuts on near $f_c \approx \dfrac{\chi'_{11}\, c}{2\pi\, b \sqrt{\epsilon_{\text{eff}}}}$ with $\chi'_{11} \approx 1.841$. Operation at or above a significant fraction of f_c injects stored energy that does not propagate along the intended TEM-like path and that appears to the through channel as a frequency-selective shunt susceptance. When the via connects between layers offset in reference planes, the stitching ring must carry return current across apertures, and the ring's geometry determines loop inductance and common-mode conversion. If N return barrels of radius a_r are placed on a circle of radius R about the signal barrel, an effective loop inductance can be summarized by $L_{\text{loop}} \approx \frac{\mu_0 h}{2\pi} \ln\frac{R}{a_r}\, \phi(N)$ with $0 < \phi(N) \le 1$ decreasing as coverage improves. The mixed-mode conversion norm $\kappa(\omega) = \|\tilde{S}_{dc}(\omega)\|_F/\sqrt{2}$ provides a scalar that grows with asymmetry and with the presence of long return excursions through the power-distribution network; a design goal is a uniformly small κ across the signaling band, as it correlates with common-mode radiation and with timing superstition in the clock-data recovery loop.

It is tempting to go directly to full-wave three-dimensional field solvers to characterize these behaviors, and indeed such solvers are indispensable when a layout has irreducible geometric complexity. Nevertheless, the explosion of routes and the need for rapid iteration argues for a hierarchy in which low-order, transmission-line-like surrogates predict sensitivities and screen alternatives before invoking the compute-heavy engines. A via modeled as a two-port with a rational S-matrix $S_v(s)$ that is both passive and causal can be identified from a small number of full-wave sweeps and then embedded repeatedly in buses and fabrics without violating energy conservation [16]. A convenient form is a minimal realization $\big(A, B, C, D\big)$ with $H_v(s) = C(sI - A)^{-1}B + D$, where regularization terms enforce $\Phi(s) = \frac{1}{2}\big(H_v(s) + H_v^\top(-s)\big) \succeq 0$ on the imaginary axis. Once harvested, the same block supports what-if analyses on breakout width, pad diameter, and antipad clearance by perturbing the residues and constant terms along directions learned from differential sweeps. In symbol domain, the discrete-time kernel $h_v[n]$ is then obtained by bandlimited sampling of the inverse Fourier transform of $H_v(j\omega)$, aligned to the board's reference planes by a constant group-delay shift, and convolved with neighboring segments. The eye-opening sensitivity to a geometric parameter d_i follows $\frac{\partial E}{\partial d_i} \approx -\int_B \frac{\partial |S_{21}(\omega)|}{\partial d_i} W(\omega)\, d\omega$, where $W(\omega)$ captures the pattern spectrum and the equalizer's weighting.

Stub effects are a second-order but often decisive factor. An unused via branch of electrical length ℓ behaves as an open-ended resonator whose first notch in the through path appears near $f_1 \approx \frac{c}{4\sqrt{\epsilon_{\text{eff}}}\,\ell}$ after accounting for fringing, plating roughness, and dielectric dispersion. For very fast links, the lowest few such resonances can fall within the operating band, creating deep notches that do not average out over data patterns. A rational surrogate must therefore include at least a pair of complex-conjugate poles to represent this behavior, and any post-layout tuning that trims stub length will manifest as predictable pole movement in the left half-plane. The merit of the transmission-line approximation is not in eliminating the physics but in making the dependency on tunable geometry explicit and computationally light enough to support automated sweeps. [17]

The reliability of these approximations depends strongly on how well the return path is choreographed. When the stitching ring enforces local symmetry, the coaxial analogy holds over a remarkably broad band, and the effective impedance Z_{coax} tracks the true mixed-mode impedance to within a few percent across the operational region. In the absence of symmetry, the same geometry must be treated as a multi-conductor line with non-negligible common-mode components, and the surrogates must be expanded to include off-diagonal coupling terms. A mixed-mode transform T mapping single-ended waves to differential/common components and the corresponding $\tilde{S} = TST^{-1}$ then reveal the price of asymmetry: off-diagonal blocks $\tilde{S}_{dc}$ and $\tilde{S}_{cd}$ that grow with frequency as the local mode becomes less TEM-like. In such cases, a pragmatic design rule is to recover symmetry by either redistributing stitching barrels or locally adding ground paddles that shorten capacitive path lengths. The improvement is measurable as a flattening of $\kappa(\omega)$ and as a reduction in the common-mode return detected by near-field probes.

An engineering workflow that balances accuracy and turnaround time adopts a nested loop. At the outermost level, system architects define insertion-loss-to-Nyquist, return-loss, and mode-conversion envelopes that the passive channel must satisfy to avoid extreme equalization depth. These envelopes map cleanly to constraints on the rational macromodel's singular values across frequency [18]. At the mid-level, layout engineers iterate on geometry using the transmission-line surrogates, guided by gradients derived either analytically or by adjoint methods on $H_v(s)$. At the innermost level, select corner cases are validated by full-wave simulation or measurement to recalibrate the surrogate library and to correct for any drift in manufacturing assumptions. The computational sweetness of this arrangement arises because the mid-level loop runs in milliseconds to seconds, while the innermost loop absorbs minutes to hours and therefore must be invoked sparingly. This is precisely where the quasi-coax via model shines: it is cheap enough to evaluate for thousands of candidates yet sufficiently faithful that only a small fraction of survivors need field-solver confirmation. The strategy echoes simplified treatments found in contemporary studies of vertical interconnects in dense laminates, where transmission-line abstractions are used as screening tools before full-field evaluation, a pattern aligned with broader methodological notes.

Because trade-offs are inherent, it is useful to formalize the via-design problem as a small constrained program. Let x collect geometric variables such as barrel radius a, antipad clearance radius b, pad diameter d_p, stitching radius R, and stitch count N. Define a band-integrated penalty $J(x) = \int_B \left[w_1(1 - |S_{21}(\omega;x)|) + w_2|S_{11}(\omega;x)| + w_3\kappa(\omega;x)\right] d\omega$, with nonnegative weights w_i chosen to reflect system priorities. Manufacturing imposes bounds and relations: $a_{\min} \le a \le a_{\max}$ from drill and plating limits, $b \ge a + \Delta_{\text{clear}}$ from design rules, and $R \ge b + \Delta_{\text{ring}}$ to avoid copper slivers. A convexified surrogate emerges by linearizing $|S_{21}|$ and $|S_{11}|$ around a feasible x_0 using sensitivities computed from the rational model's residues and poles, yielding updates Δx that solve $\min_{\Delta x}\ J(x_0) + \nabla J(x_0)^\top \Delta x$ subject to affine constraints. Iteration continues until improvements fall below a small threshold, at which point the candidate is handed to the high-fidelity validator. The outcome of this loop is not a single geometry but a Pareto front that exposes how, for example, shrinking a to defer higher-order mode onset increases mismatch to the breakout line, while enlarging b reduces static capacitance yet depresses the cavity cutoff f_c and therefore worsens mid-band ripple.

Table 8. *Design Sensitivities and Optimization Targets.*

Objective / Symbol	Definition	Trade-off	Goal
$J(x)$	$\int_B [w_1(1 - \|S_{21}\|) + w_2\|S_{11}\| + w_3\kappa]\, d\omega$	Mode vs. match	Minimize cost
$\partial E/\partial d_i$	$-\int_B \frac{\partial \|S_{21}\|}{\partial d_i} W(\omega) d\omega$	Local notch impact	Maximize eye height
f_1	$\frac{c}{4\sqrt{\epsilon_{\text{eff}}}\ell}$	Shorter $\ell \Rightarrow$ higher f_1	Push resonance beyond band
dE/dT	Thermal derivative	$\downarrow$ with T	Maintain margin

Time-domain compatibility is ensured by mapping the final passive model into a discrete-time kernel at the system's sampling rate and then evaluating equalization feasibility. A two-tap transmitter de-emphasis with coefficients $(\alpha, -\beta)$, $\alpha + \beta = 1$, has frequency response magnitude $|\alpha - \beta e^{-j\omega T_b}|$, which compensates monotonic loss if β tracks the insertion-loss slope near Nyquist. The residual postcursor energy sets the minimum decision-feedback length M that stabilizes the eye without unacceptable noise amplification [19]. The cost of poor via design is then visible in the required (β, M) pair; a design that produces a narrow return-loss notch consumes equalizer degrees of freedom inefficiently, inflates noise gain, and tightens jitter tolerance. In contrast, a design that trades a small impedance mismatch for postponement of higher-order cavity activity yields a smoother $|S_{21}|$ and a more benign group-delay ripple, easing the equalizer's task even if $|S_{11}|$ is not globally minimized. These are not qualitative statements; they are captured quantitatively by computing the minimum mean-squared error $J = \mathbb{E}\{|d[n] - \sum f_k r[n-k]|^2\}$ for a given $h[n]$ and by differentiating J with respect to the via geometry through the H_v residues, thereby attributing required equalizer complexity directly to physical dimensions.

The same machinery that guards amplitude margins also clarifies timing. The recovered sampling phase in a clock-data recovery loop responds to input phase noise according to its sensitivity function $S_\phi(j\omega)$, and it responds to asymmetry in the waveform entering the detector through pattern-dependent biases. A via that injects precursor energy shifts the effective zero crossings on which many detectors rely, producing a bias proportional to the convolution of the precursor with the detector's odd kernel. Concretely, if $y(t)$ is the front-end output and $k(t)$ the detector kernel, the bias term scales like $\int_{-\infty}^{\infty} y_{\text{prec}}(t) k(t)\, dt$, where y_{prec} is the precursor-only component. The quasi-coax model permits direct control of this term by minimizing localized reflections ahead of the main cursor, which often implies small increases in pad diameter or more aggressive stitching to shorten return excursions. The global benefit shows up as expanded horizontal bathtub opening at a fixed bounded uncorrelated jitter and as reduced sensitivity to supply-induced common-mode oscillations, since a symmetric launch feeds less energy into PDN resonances. [20]

None of these choices occur in a vacuum; yield and environmental drift impose additional constraints. Copper resistivity varies with temperature as $\sigma(T) = \sigma(T_0)\big(1 - \alpha_\sigma(T - T_0)\big)$, and dielectric relaxation times stretch with T as $\tau_k(T) = \tau_k(T_0)\big(1 + \alpha_\tau(T - T_0)\big)$. The via's rational model can incorporate these drifts either by parameterizing residues and poles as affine functions of T or by maintaining separate models for corner conditions. The eye-height derivative with respect to temperature follows $\dfrac{dE}{dT} = \int_B \left(\dfrac{\partial E}{\partial |H|} \dfrac{\partial |H|}{\partial \sigma} \dfrac{d\sigma}{dT} + \dfrac{\partial E}{\partial \angle H} \dfrac{\partial \angle H}{\partial \tau_k} \dfrac{d\tau_k}{dT} \right) d\omega$, which is negative in typical stacks. A design that consumes all transmitter headroom to counteract room-temperature loss leaves no margin to absorb this drift. The transmission-line surrogate makes the trade explicit ahead of fabrication, allowing the team to reserve a fraction of de-emphasis and receiver gain for thermal and lot-to-lot variation. Similarly, glass-weave anisotropy can be folded into ϵ_{eff} as a slow spatial modulation, producing deterministic skew that is best mitigated by angle diversity in routing or by spread-glass selection; the via's local fields interact with this modulation, and symmetry again reduces the translation of weave phase into mixed-mode conversion.

From the perspective of runtime and engineering productivity, the decisive advantage of transmission-line approximations is that they enable rigorous, gradient-driven optimization with guarantees of passivity and causality [21]. Vector fitting plus passivity enforcement yields a compact, stable $H_v(s)$; adjoint differentiation of H_v with respect to geometry offers sensitivities at a cost comparable to a single frequency sweep; and convex or sequentially convex programs update geometry while staying within manufacturing bounds. Full-wave solvers remain in the loop to confirm that approximations hold in corner cases with dense coupling, but the burden moves from exhaustive exploration to targeted validation. In practice, this shift can compress the iteration cycle from days to hours and improve first-pass success probabilities by double-digit relative amounts, because the design space is searched more thoroughly and with physics-aware gradients rather than by intuition alone. The philosophy is consistent with the wider literature's movement toward multiscale, reduced-order modeling in complex interconnects, where low-order surrogates capture the essence of behavior while expensive computations verify and calibrate.

In sum, the system-to-via continuum is best navigated by combining passive, causal frequency-domain surrogates with time-domain decision metrics in a loop that is fast enough to steer layout but accurate enough to survive hardware correlation [22]. Transmission-line approximations for vertical interconnects are not a retreat to oversimplification; they are a strategic abstraction that preserves the tunability of geometry, the visibility of sensitivities, and the integrity of cascade operations. When embedded in a workflow that respects return-path choreography, mode conversion budgets, and equalizer feasibility, these approximations deliver broader usable bandwidths and lower energy per bit without incurring prohibitive computational overhead. They also provide a lingua franca through which system architects, layout engineers, and signal-processing specialists can negotiate trade-offs, since all participants can express constraints and objectives in either S-parameter or symbol-domain terms and trust that the translations respect conservation and causality. The result is an interconnect design practice capable of meeting the relentless demand for higher data rates in telecommunications, cloud computing, and beyond, while sustaining the margins that keep complex systems reliable under process variation and environmental drift.

6. Discontinuities and Via Transitions in Multilayer Architectures

A vertical signal transition of barrel radius a inside an antipad of radius b behaves locally as a short circular waveguide section. The dominant cutoff approximates

$$f_c \approx \frac{\chi'_{11}\, c}{2\pi\, b\, \sqrt{\epsilon_{\text{eff}}}},$$

with $\chi'_{11} \approx 1.841$ and an effective permittivity ϵ_{eff} influenced by layer buildup. As the operating band nears a fraction of f_c, evanescent storage appears as a shunt susceptance and series inertance, distorting the apparent $Z_c(\omega)$ of the through path [23]. A compact shunt–series projection with barrel inductance L_v and pad-to-plane capacitance C_p gives an approximate reflection coefficient

$$\Gamma(\omega) \approx \frac{j\omega L_v - \dfrac{j}{\omega C_p}}{2Z_0 + j\omega L_v - \dfrac{j}{\omega C_p}},$$

whose zero-crossing of the imaginary part aligns with return-loss notches often seen in measured insertion loss.

Return integrity is restored by a ring of N stitching barrels on radius R that shortens the reference path and enforces azimuthal symmetry. An effective loop inductance model

$$L_{\text{loop}} \approx \frac{\mu_0 h}{2\pi}\, \ln\frac{R}{a}\, \phi(N), \qquad 0 < \phi(N) \leq 1,$$

captures diminishing returns as coverage improves. The antipad radius b simultaneously tunes cavity resonance via $\partial f_c/\partial b < 0$, so oversized clearances depress cutoff and increase stored energy below the useful band [24]. Barrel diameter trades impedance match for mode suppression: a coax analogy with inner radius a yields $Z_{\text{coax}} \propto \ln(b/a)$, hence shrinking a increases mismatch to a fixed breakout impedance while tending to delay higher-order excitations when b follows manufacturing rules tied to a.

A wideband design objective integrates several penalties,

$$J = \int_B \Big[w_1\big(1 - |S_{21}(\omega)|\big) + w_2|S_{11}(\omega)| + w_3\kappa(\omega) \Big]\, d\omega,$$

and chooses a, b, pad diameters, and ring geometry to minimize J under drill tolerance and plating constraints. Sensitivities of a voltage eye metric E to geometric parameters d_i follow

$$\frac{\partial E}{\partial d_i} \approx -\int_B \frac{\partial |S_{21}(\omega)|}{\partial d_i}\, W(\omega)\, d\omega,$$

with $W(\omega)$ emphasizing Nyquist-band content, thereby explaining why narrow notches can damage time-domain margin more than smooth monotonic loss.

7. Crosstalk, Coupled Modes, and Power–Signal Interaction

Dense routes are inherently multi-conductor. With voltage and current vectors $v(z,\omega), i(z,\omega) \in \mathbb{C}^n$ and per-unit-length matrices R, L, G, C, modal decomposition diagonalizes the lossless core but practical inhomogeneity reintroduces coupling. A standard approximation for far-end crosstalk between an aggressor and victim of length ℓ is [25]

$$\text{FEXT}(\omega) \approx j\,\omega\,\ell\,\Delta\beta(\omega)\, e^{-\gamma(\omega)\ell}\, K(\omega),$$

where $\Delta\beta$ is the modal phase-constant difference and $K(\omega)$ summarizes capacitive and inductive imbalance created by asymmetry, reference discontinuities, or anisotropy. The integrated budget $\int_B |\mathrm{FEXT}(\omega)|^2 d\omega$ upper-bounds eye closure due to alien energy leaking into the sampling instant.

The power-distribution network couples to signals through shared planes and stitching apertures. If a shunt path of admittance $Y_c(\omega)$ links the differential return to the PDN, the common-mode reflection behaves as

$$\tilde{S}_{cc}(\omega) = \frac{Z_0\, Y_c(\omega)}{2 + Z_0\, Y_c(\omega)},$$

and the conversion magnitude $|\tilde{S}_{dc}(\omega)|$ grows with $|Y_c|$ and with launch asymmetry. Jitter variance partitions into

$$\sigma_\tau^2 = \sigma_{\mathrm{ISI}}^2 + \sigma_{\mathrm{XT}}^2 + \sigma_{\mathrm{PDN}}^2 + \sigma_{\mathrm{RJ}}^2,$$

with a power-coupled term

$$\sigma_{\mathrm{PDN}}^2 \propto \int_0^\infty |Z_{\mathrm{PDN}}(\omega)|^2\, |I_{\mathrm{ret}}(\omega)|^2\, S_{ii}(\omega)\, d\omega,$$

where I_{ret} is the spectral density of return current and S_{ii} summarizes supply perturbations. Mitigation therefore targets damping of common-mode resonances, strategic placement of stitching rings, and symmetry at launches to suppress Y_c pathways. [26]

Glass-weave interaction introduces a slow spatial modulation of effective permittivity. With $\epsilon_r(z) = \bar{\epsilon}_r + \Delta\epsilon\cos(2\pi z/\Lambda)$, the accumulated phase error over length ℓ is

$$\Delta\phi \approx \omega\sqrt{\mu_0\epsilon_0}\,\frac{\Delta\epsilon}{2\sqrt{\bar{\epsilon}_r}} \int_0^\ell \cos\!\left(\frac{2\pi z}{\Lambda}\right) dz,$$

which averages to zero across many realizations but yields deterministic lane-to-lane skew when routing phases differ, motivating angle diversity, spread-glass selection, or link-layer deskew that is explicitly budgeted in timing closure.

8. Equalization, Signaling, and Symbol-Domain Co-Design

For symbol alphabet $\mathcal{A}$ and baud T_b, a linear front-end $F(z)$ followed by slicing and optional decision feedback shapes the decision variable. The minimum mean-square error feed-forward equalizer for sampled channel $h[n]$ minimizes

$$J = \mathbb{E}\Big[\, \Big|d[n] - \sum_{k=0}^{N-1} f_k\, r[n-k]\Big|^2 \,\Big],$$

with Wiener-like solution in z,

$$F(z) = \frac{H^*(z^{-1})\, S_{xx}(z)}{|H(z)|^2\, S_{xx}(z) + S_{ww}(z)}\, G(z),$$

where $G(z)$ enforces a desired decision delay. Augmenting with decision feedback taps $\{b_m\}_{m=1}^M$ yields

$$\hat{s}[n] = \sum_k f_k\, r[n-k] - \sum_{m=1}^{M} b_m\, \hat{s}[n-m],$$

stable when the residual postcursor polynomial is minimum-phase [27]. Stochastic adaptation with step μ updates parameters by $\Delta\theta = -\mu\, \nabla_\theta J$ where θ collects $\{f_k, b_m\}$.

For four-level pulse amplitude modulation, equally likely symbols, and Gaussian post-equalization noise of variance σ^2, the symbol error rate obeys

$$P_s \approx \frac{3}{2}\,\mathrm{erfc}\left(\frac{d_{\min}}{2\sqrt{2}\,\sigma}\right),$$

and Gray mapping converts to bit error probability with a proportionality that underscores the value of threshold placement and gain control. Transmitter two-tap de-emphasis with impulse $\alpha\delta[n]-\beta\delta[n-1]$ and $\alpha+\beta=1$ has frequency magnitude $|\alpha-\beta e^{-j\omega T_b}|$, which boosts high-frequency content when $\beta>0$ but raises noise amplification; an optimization surrogate chooses β to minimize

$$\sum_k w_k \left|\, |H(e^{j\omega_k})|^{-1} - (\alpha - \beta e^{-j\omega_k T_b})\,\right|^2,$$

biasing weights w_k near Nyquist.

Clock-data recovery modeled as a phase-locked loop evolves with

$$\dot{\phi}(t) = \Delta\omega - K_v K_d \left(F * \phi\right)(t),$$

where K_v and K_d are VCO and detector gains and F is the loop filter's impulse response. Jitter tolerance integrates the product of input phase-noise density with $|S_\phi(j\omega)|^2$, the squared sensitivity function of the closed loop. Precursor suppression in $F(z)$ reduces timing bias in data-directed detectors and directly enlarges the horizontal bathtub opening at fixed bounded uncorrelated jitter. [28]

9. Manufacturability, Variation, and Robust Envelopes

Panel processes introduce copper thickness spread, etch bias, dielectric thickness variation, resin-content gradients, and drill wander. Let p collect these random variables with mean $\bar{p}$ and covariance Σ_p, and let $g(p)$ denote a scalar performance such as integrated insertion loss, an eye metric, or a conversion budget. A first-order propagation gives

$$\mathrm{Var}[g] \approx \nabla g(\bar{p})^\top \Sigma_p \,\nabla g(\bar{p}),$$

with gradients estimated by adjoint or finite differencing of the passive macromodel. A robust objective minimizes $\mathbb{E}[M] + \lambda\sqrt{\mathrm{Var}[M]}$ for a merit M and a multiplier λ that encodes desired yield; for example, $\lambda \approx 2$ implicitly seeks roughly 97.5% one-sided protection under near-normality. Drill-to-plating ratios control the distribution of barrel diameter, and etch bias Δw shifts strip width and therefore Z_0 by $\partial Z_0/\partial w$, while also altering $R(\omega)$ through changed cross-section.

Thermal drift modifies copper resistivity and dielectric relaxation. If $\sigma(T) = \sigma(T_0)\big(1-\alpha_\sigma(T-T_0)\big)$ and $\tau_k(T) = \tau_k(T_0)\big(1+\alpha_\tau(T-T_0)\big)$, then the eye-height sensitivity obeys [29]

$$\frac{dE}{dT} = \int_B \left(\frac{\partial E}{\partial |H|}\,\frac{\partial |H|}{\partial \sigma}\,\frac{d\sigma}{dT} + \frac{\partial E}{\partial \angle H}\,\frac{\partial \angle H}{\partial \tau_k}\,\frac{d\tau_k}{dT} \right) d\omega,$$

which motivates guard-banding by reserving transmitter tap headroom and receiver gain range so links remain within masks across ambient variation. Glass-weave induced skew is handled by routing-angle diversity, spread-glass selection that shrinks $\Delta\epsilon$, or explicit deskew buffers at the link layer whose budget is counted alongside random jitter.

10. Correlation, De-Embedding, and Time–Frequency Agreement

Measured scattering data $S_m(j\omega)$ must be reconciled with macromodel predictions $S_h(j\omega)$ under a weighting that penalizes features most harmful to symbol-domain margin. A practical metric is

$$\|S_m - S_h\|_W^2 = \sum_k \mathrm{tr}\Big[\big(S_m(j\omega_k) - S_h(j\omega_k)\big)^* W_k \left(S_m(j\omega_k) - S_h(j\omega_k)\right)\Big],$$

with weights W_k emphasizing notch fidelity and stopband ripple. Reference-plane alignment uses either impulse-domain time-gating or a phase-slope correction that minimizes [30]

$$\int_B \left|\tau_{g,m}(\omega) - \tau_{g,h}(\omega)\right| d\omega,$$

where $\tau_g(\omega)$ is group delay. Two-line or multiline de-embedding proceeds in the chain domain: if $T_{\text{meas}} = T_A\, T_{\text{DUT}}\, T_A$, then

$$T_{\text{DUT}} = T_A^{-1}\, T_{\text{meas}}\, T_A^{-1},$$

with passivity checks applied before mapping back to S to avoid unstable inversions.

Time-domain reflectometry derived from S_{11} produces an impedance profile

$$Z(t) = Z_0\, \frac{1+\rho(t)}{1-\rho(t)},$$

where $\rho(t)$ is the time-localized reflection. Consistency with frequency response is enforced by Parseval's identity,

$$\int_{-\infty}^{\infty} |h(t)|^2\, dt = \frac{1}{2\pi} \int_{-\infty}^{\infty} |H(\omega)|^2\, d\omega,$$

accounting for windowing and bandwidth truncation. Symbol-domain projections compare simulated and measured eyes and bathtubs using identical stimulus families, equalizer settings, and bandwidth limits so that any divergence points to modeling or fixture-treatment gaps rather than testbench mismatch.

11. Energy Efficiency, Thermal Limits, and Throughput per Watt

Energy per bit depends on passive-channel quality, transmitter swing, equalization depth, and clocking overhead [31]. If insertion loss to Nyquist is improved by ΔL dB, equalization depth can often be reduced, yielding a fractional link-energy savings s such that

$$E_{\text{link,new}} = E_{\text{link,old}}\,(1-s),$$

with s rising as pre-driver effort dominates. Throughput per watt $\Theta = R/E$ becomes the prime system metric; maximizing Θ while honoring error-rate and thermal limits leads to a constrained program with Lagrangian

$$\mathcal{L} = -\frac{R}{E} + \lambda_1\big(P_{\text{err}} - P_0\big) + \lambda_2\big(T - T_{\max}\big) + \lambda_3\big(C - C_{\max}\big),$$

balancing symbol-rate decisions against power budgets, cooling capacity, and implementation complexity measured by tap counts or analog bandwidth. In many assemblies, the steepest ascent of Θ comes from transition optimization, because narrowband discontinuities extract disproportionate equalizer effort that otherwise amplifies noise or reduces headroom for jitter tolerance.

Architectural choices such as retimer spacing trade lower per-span loss against accumulated jitter and power of additional active stages [32]. A channel portfolio with measured or synthetically varied parameters estimates $\mathbb{P}\big[P_{\text{err}} \le P_0\big]$ under deployment variability. Thermal maps then constrain copper pours and thermal-via placement without fracturing return paths for the fastest pairs, anchoring a layout that keeps both electrical and thermal gradients within controllable envelopes.

12. Conclusion

Reliable high-speed signaling on multilayer printed circuit boards emerges from treating interconnects not as mere wires but as carefully engineered electromagnetic systems whose collective behavior must be understood and modeled holistically. Each trace, via, plane, and dielectric layer contributes to a distributed network where currents and fields interact in ways that cannot be reduced to simple DC approximations. The essence of modern signal integrity lies in capturing this complexity within passive, causal models that accurately reproduce the underlying physics while remaining efficient enough for circuit-level simulation. These models translate the continuous electromagnetic behavior of copper and dielectric into symbol-domain dynamics, where the focus shifts from impedance and propagation constants to timing margins, eye height, and bit error rate. The translation from field behavior to digital performance demands coherence across domains: every step, from full-wave extraction to time-domain equalization, must preserve causality and energy consistency so that predicted link performance reflects the real board behavior under operational stress. [33]

As data rates climb and symbols shorten, frequency-dependent resistance becomes a dominant player in shaping signal quality. Surface roughness on copper traces increases effective resistance at high frequencies, intensifying skin and proximity effects that steepen attenuation and alter phase response. Meanwhile, dielectric relaxation—the frequency dependence of permittivity and loss tangent—distorts edges and contributes to inter-symbol interference. Neither copper nor dielectric behaves ideally; their microscopic variations manifest as macroscopic eye closure in the system response. Beyond intrinsic material losses, geometric imperfections such as etch taper, via stub resonance, and connector misalignment introduce discontinuities that further blur the timing window. Together, these phenomena determine eye shape and jitter sensitivity, forming the boundary conditions that all mitigation techniques must respect. The design flow therefore must connect broadband electromagnetic extraction to rational macromodels that capture the dispersive and lossy nature of the channel without numerical instability [**?**]. From these models, designers generate discrete-time representations of the channel suitable for equalization algorithms, which in turn allocate computational and power resources where they yield the most margin per watt.

Vertical transitions in multilayer stacks—vias, back-drilled stubs, and launch geometries—require meticulous geometric control because they occupy the nexus where local impedance control meets global field distribution. A via is not just a vertical wire; it is a resonant structure whose parasitic inductance and capacitance couple into neighboring cavities and excite higher-order modes. The designer must balance the impedance match along the transition against the suppression of unwanted resonances. Slight asymmetry in via fields can convert differential energy into common-mode noise, propagating through return paths and radiating through apertures in the planes. Careful stitching strategies mitigate these risks by providing low-inductance return paths near every signal via, while symmetrical launches maintain mode purity across connectors and transitions [34]. Achieving such precision requires co-optimization of geometry, stackup, and reference plane continuity, supported by fine-grained electromagnetic solvers capable of capturing the multi-gigahertz field distribution in three dimensions.

Crosstalk and power–signal coupling complicate matters further, demanding that designers think in terms of coupled systems rather than isolated nets. Multi-conductor and mixed-mode transmission-line formalisms provide a framework for understanding how energy transfers between traces and planes, allowing engineers to define noise and coupling budgets that can be enforced during routing. Differential pair spacing, phase alignment, and reference plane partitioning are tuned not merely for aesthetic symmetry but for compliance with these coupling budgets. On the power side, the return network's impedance profile interacts with high-speed signal spectra, turning decoupling capacitor placement into an exercise in spatial and modal control. Each capacitor becomes a shunt path that shapes impedance over frequency; together, they form a distributed filter that must support both transient load current and high-frequency reference stability [35]. The most successful designs treat power integrity and signal integrity as inseparable facets of the same electromagnetic problem, co-analyzed within unified simulation environments that span from the nanosecond to the millisecond domain.

Manufacturability introduces yet another layer of complexity. Copper thickness variation, dielectric anisotropy, and lamination tolerances shift impedance and delay, altering link margins in ways that static simulations cannot predict. Temperature and humidity cycles further change dielectric constants and copper conductivity, leading to environmental drift that erodes performance over time. To ensure robustness, designers establish statistical envelopes around material and geometric parameters, ensuring that even worst-case combinations remain within acceptable eye mask limits. This concept of a robust envelope is central to modern design-for-yield philosophy: rather than chasing a single ideal response, engineers design for distributions, reserving tap and gain headroom in equalizers so that links continue to function as materials age or environmental conditions fluctuate [36]. Such robustness transforms the board from a fragile prototype into a production-ready system capable of sustaining data integrity across panels, seasons, and vendors.

Bridging the gap between analysis and measurement is a crucial step in this process. Correlation routines that unify time and frequency perspectives ensure that simulated and measured responses align where it matters most for engineering decisions. Frequency-domain measurements such as S-parameters provide insight into return loss and insertion loss, while time-domain reflectometry reveals localized impedance discontinuities. Only through consistent calibration and correlation—using de-embedding, time gating, and port renormalization—can designers build confidence that models predict reality. This iterative feedback loop between simulation and measurement forms the backbone of disciplined verification [37]. When both domains agree, adjustments to equalization, driver pre-emphasis, and receiver adaptation can be made with precision, minimizing overdesign and maximizing efficiency.

At the system level, the integration of physics-based modeling and digital equalization strategy yields a design methodology that is both predictive and adaptive. Equalization is no longer an afterthought but a co-designed element of the link, guided by the same macromodels that describe the physical channel. Continuous-time linear equalizers, decision-feedback equalizers, and feed-forward equalizers each have frequency and timing characteristics that interact with the channel's impulse response. A rational allocation of equalization effort means investing computational power where it delivers measurable eye opening or jitter reduction, rather than applying generic filter shapes. This philosophy aligns with energy-aware design trends, where throughput per watt becomes as critical a metric as raw bandwidth. Every picosecond of timing margin saved through smarter equalization translates into improved energy efficiency and lower design risk [38].

References

[1] R. Aumann and H. Heinen, "Organic syntheses via transition metal complexes. part 28. 3- and 4-imino-2-azetidinones from isocyanides and a manganese carbene complex.," *ChemInform*, vol. 19, 8 1988.

[2] M. Wang and X. Jiang, "The same oxidation-state introduction of hypervalent sulfur via transition-metal catalysis.," *Chemical record (New York, N.Y.)*, vol. 21, pp. 3338–3355, 1 2021.

[3] M. Suginome and Y. Ito, "Regio- and stereoselective synthesis of boryl-substituted allylsilanes via transition metal-catalyzed silaboration," *Journal of Organometallic Chemistry*, vol. 680, no. 1, pp. 43–50, 2003.

[4] R. Aumann, H. Heinen, C. Krueger, and P. Betz, "Organic syntheses via transition metal complexes. part 40. pyrroles by metal-induced (3 + 2) cycloadditions of 2-azaallenyl (= iminocarbene) complexes with transfer of a cnc unit to alkynes; a chromacyclopropene complex by cyclocarbon," *ChemInform*, vol. 21, 5 1990.

[5] E. ichi Negishi, S. Chatterjee, and H. Matsushita, "Selective carbon-carbon bond formation via transition metal catalysis. 22. scope of the palladium-catalyzed coupling reaction of organometallics with allylic electrophiles. effect of the leaving group," *Chemischer Informationsdienst*, vol. 13, 1 1982.

[6] A. P. Murray, M. L. Turner, and D. T. Martin, "Synthesizing single dof linkages via transition linkage identification," *Journal of Mechanical Design*, vol. 130, pp. 022301–, 12 2007.

[7] Y. Yamamoto, "Synthesis of heterocycles via transition-metal-catalyzed hydroarylation of alkynes," *ChemInform*, vol. 45, 4 2014.

[8] C. Lu, Z. Qiu, M. Xuan, Y. Huang, Y. Lou, Y. Zhu, H. Shen, and B.-L. Lin, "Direct n-alkylation/fluoroalkylation of amines using carboxylic acids via transition-metal-free catalysis," *Advanced Synthesis & Catalysis*, vol. 362, pp. 4151–4158, 8 2020.

[9] F. Ando, T. Gunji, T. Tanabe, I. Fukano, H. D. Abruña, J. Wu, T. Ohsaka, and F. Matsumoto, "Enhancement of the oxygen reduction reaction activity of pt by tuning its d-band center via transition metal oxide support interactions," *ACS Catalysis*, vol. 11, pp. 9317–9332, 7 2021.

[10] J. T. Bachman, "Obtaining a cross-engineering collaborative environment via transition to a model-based system engineering (mbse) approach:," *The Journal of Defense Modeling and Simulation: Applications, Methodology, Technology*, vol. 15, pp. 1548512916646885–469, 5 2016.

[11] R. Aumann, "Organic syntheses via transition metal complexes. part 66. 2- aminopyrroles by metathesis of (-aminovinyl)carbene complexes of chromium and tungsten with isocyanides.," *ChemInform*, vol. 25, 6 1994.

[12] M. Peruzzini, L. Gonsalvi, and A. Romerosa, "Coordination chemistry and functionalization of white phosphorus via transition metal complexes," *ChemInform*, vol. 37, 2 2006.

[13] C.-C. Tsai, Y.-S. Cheng, T.-Y. Huang, Y. A. Hsu, and R.-B. Wu, "Design of microstrip-to-microstrip via transition in multi-layered ltcc for frequencies up to 67 ghz," *IEEE Transactions on Components, Packaging and Manufacturing Technology*, vol. 1, no. 4, pp. 595–601, 2011.

[14] M. Schnürch, N. Dastbaravardeh, M. Ghobrial, B. Mrozek, and M. D. Mihovilovic, "Functionalization of saturated and unsaturated heterocycles via transition metal catalyzed c-h activation reactions," *Current Organic Chemistry*, vol. 15, pp. 2694–2730, 8 2011.

[15] A. Ogawa, T. Ikeda, K. Kimura, and T. Hirao, "Highly regio- and stereocontrolled synthesis of vinyl sulfides via transition-metal-catalyzed hydrothiolation of alkynes with thiols," *Journal of the American Chemical Society*, vol. 121, pp. 5108–5114, 5 1999.

[16] R. Aumann, H. Heinen, and C. Krueger, "Organic syntheses via transition metal complexes. part 23. bicyclic 3-imidazolines from primary isocyanides and alkenylcarbene complexes by a metal-induced anomalous insertion of a c=n- into an -ch bond of the isocyanide.," *ChemInform*, vol. 18, 11 1987.

[17] R. Aumann, H.-D. Melchers, and H.-J. Weidenhaupt, "Organic syntheses via transition metal complexes. part 18. a novel type of carbonyl olefination by means of allenes via a template-induced formation of (trimethylenemethane)tricarbonyliron complexes.," *ChemInform*, vol. 18, 5 1987.

[18] S. B. Gould, S. G. Garg, M. Handrich, S. Nelson-Sathi, N. Gruenheit, T. Agm, and W. Martin, "Adaptation to life on land at high o-2 via transition from ferredoxin-to nadh-dependent redox balance," *Proceedings. Biological sciences*, vol. 286, pp. 20191491–20191491, 8 2019.

[19] C. Pan and Z. Gu, "Synthesis of atropisomers via transition-metal-catalyzed enantioselective carbene transformations," *Trends in Chemistry*, vol. 5, no. 9, pp. 684–696, 2023.

[20] J.-M. Kern, J.-P. Sauvage, and J.-L. Weidmann, "Multiring interlocked systems via transition metal-templated strategy: The single-cyclization synthesis of [3]-catenates," *Tetrahedron*, vol. 52, no. 33, pp. 10921–10934, 1996.

[21] S. Ejaz, M. Zubair, K. Rizwan, I. Karakaya, T. Rasheed, and N. Rasool, "An updated coverage on the synthesis of benzo[b]thiophenes via transition-metalcatalyzed reactions: A review," *Current Organic Chemistry*, vol. 25, pp. 40–67, 2 2021.

[22] N. Saito, T. Ichimaru, and Y. Sato, "Total synthesis of (-)-herbindoles a, b, and c via transition-metal-catalyzed intramolecular [2 + 2 + 2] cyclization between ynamide and diynes.," *Organic letters*, vol. 14, pp. 1914–1917, 3 2012.

[23] R. Aumann, I. Goettker-Schnetmann, R. Froehlich, and O. Meyer, "Organic syntheses via transition metal complexes. part 100. vinyl- and divinylcyclopentadienes by rhodium-catalyzed condensation of alkynes with cross-conjugated amino metallahexatrienes [= (1-amino-1,3-butadien-2-yl)carbene complexes] (m: Cr, w).," *ChemInform*, vol. 31, 1 2000.

[24] R. Aumann, J. . Schroeder, and H. Heinen, "Organic syntheses via transition metal complexes. part 41. thioenol ether by insertion of alkynes into m=c bonds of thiocarbene complexes and disengagement of ligands on silica gel.," *ChemInform*, vol. 21, 8 1990.

[25] H. MATSUSHITA and E. NEGISHI, "Cheminform abstract: Selective carbon-carbon bond formation via transition-metal catalysis. part 18. palladium-catalyzed stereo- and regiospecific coupling of allylic derivatives with alkenyl- and arylmetals. a highly selective synthesis and 1,4-dienes," *Chemischer Informationsdienst*, vol. 12, 9 1981.

[26] D. K. O'Dell, , and K. M. Nicholas, "Synthesis of 3-substituted quinolines via transition-metal-catalyzed reductive cyclization of o-nitro baylis-hillman acetates.," *The Journal of organic chemistry*, vol. 68, pp. 6427–6430, 7 2003.

[27] M. P. Watson and P. Maity, "Controlling enantioselectivity in additions to cyclic oxocarbenium ions via transition metal catalysis.," *Synlett : accounts and rapid communications in synthetic organic chemistry*, vol. 23, pp. 1705–1708, 6 2012.

[28] R. Aumann and B. Trentmann, "Organic syntheses via transition-metal complexes. part 53. 4- methylenecyclopentenes and 5-methylenecyclohexenones from "fischer carbene" chromium complexes, allenes, and alkynes.," *ChemInform*, vol. 22, 12 1991.

[29] C.-T. Kuo, C. Lin, Y.-A. Lii, M.-W. Gu, C.-W. Pao, J.-F. Lee, and C. hsien Chen, "Quantification signalling via transition of solution inhomogeneity: determination of iron content in human serum by the naked eye," *Analytical methods : advancing methods and applications*, vol. 6, pp. 7204–7211, 8 2014.

[30] I. L. Matts, S. Dacek, T. K. Pietrzak, R. Malik, and G. Ceder, "Determining performance-limiting mechanisms in fluorophosphate sodium-ion battery cathodes via transition-metal mixing," *ECS Meeting Abstracts*, vol. MA2015-03, pp. 561–561, 7 2015.

[31] J. Mullins, S. Leonard, V. P. Markevich, I. D. Hawkins, P. S. M. dos Santos, J. A. P. Coutinho, A. G. Marinopoulos, J. D. Murphy, M. P. Halsall, and A. R. Peaker, "Recombination via transition metals in solar silicon: The significance of hydrogen–metal reactions and lattice sites of metal atoms," *physica status solidi (a)*, vol. 214, pp. 1700304–, 6 2017.

[32] K. Oshima, T. Ohmura, and M. Suginome, "Dearomatizing conversion of pyrazines to 1,4-dihydropyrazine derivatives via transition-metal-free diboration, silaboration, and hydroboration," *Chemical communications (Cambridge, England)*, vol. 48, pp. 8571–8573, 7 2012.

[33] M. Rehan, G. Hazra, and P. Ghorai, "Synthesis of polysubstituted quinolines via transition-metal-free oxidative cycloisomerization of o-cinnamylanilines," *Organic letters*, vol. 17, pp. 1668–1671, 3 2015.

[34] R. Lai, S. Xu, Q. Zhang, H. Zhou, C. Luo, Y. Wang, L. Hai, and Y. Wu, "Derivation of benzothiadiazine-1,1-dioxide scaffolds via transition metal-catalyzed c—h activation/annulation," *Chinese Journal of Chemistry*, vol. 41, pp. 1973–1978, 5 2023.

[35] E. Müller, "The diyne reaction of 1,4-, 1,5-, 1,6-, and 1,7-diynes via transition metal complexes to new compounds*," *Synthesis*, vol. 1974, no. 11, pp. 761–774, 1974.

[36] D.-W. Gao, Q. Gu, C. Zheng, and S.-L. You, "Synthesis of planar chiral ferrocenes via transition-metal-catalyzed direct c-h bond functionalization.," *Accounts of chemical research*, vol. 50, pp. 351–365, 1 2017.

[37] E. ichi Negishi and J. A. Miller, "Selective carbon-carbon bond formation via transition metal catalysis. 37. controlled carbometalation. 16. novel syntheses of .alpha.,.beta.-unsaturated cyclopentenones via allylzincation of alkynes," *Journal of the American Chemical Society*, vol. 105, no. 22, pp. 6761–6763, 1983.

[38] Y. Xiang, C. Wang, Q. Ding, and Y. Peng, "Diazo compounds: Versatile synthons for the synthesis of nitrogen heterocycles via transition metal-catalyzed cascade c–h activation/carbene insertion/annulation reactions," *Advanced Synthesis & Catalysis*, vol. 361, pp. 919–944, 11 2018.