Approximation of Worst Case Crosstalk at High Frequencies

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***Abstract*—Predicting system-level crosstalk problems early in the design process is increasingly important to today's automotive industry. Closed-form estimates of worst or nearly worst case crosstalk can be useful in system analysis, since these estimates can quickly identify problem areas that require close inspection and they show a clear relationship between design parameters and crosstalk that allows intelligent modification of the system. In this paper, simple formulas are derived to estimate worst case crosstalk between circuits for weak coupling at high frequencies (i.e., where the length of the circuits is comparable to or greater than the wavelengths of the signals of interest). Derivations are based on transmission line theory for general (inductive, capacitive, or resistive) sources and loads, assuming losses at the terminations dominate losses in the transmission lines. Formulas are verified with bench-top and in-vehicle experiments and are shown to predict maximum crosstalk levels within about 6 dB.**

***Index Terms*—Cables, coupling circuits, crosstalk, transmission line, vehicles.**

I. Introduction

CROSSTALK problems in modern automotive systems may cause failures or malfunctions that lead to both safety and reliability issues. When problems are not discovered until late in the design cycle (e.g., after a prototype has been produced), it is often too late to implement a solution that is both practical and cost effective. To maximize quality while minimizing costs, the automotive engineer must uncover, understand, and eliminate crosstalk problems early in the design process.

Several methods have been explored to determine crosstalk in cable-harness bundles [1]–[21]. Most modern methods rely on detailed numerical models [5]–[11]. Numerical methods, however, often suffer both from the complexity of the system (e.g., how does one get the information in a form that may be used by the solver?) and from the uncertainty of system parameters early in the design process. Many parameters are unknown and must be estimated or are intrinsically random in nature, like the position of a wire in the harness bundle. Because these numerical methods find the solution for only one specific configuration, iterative techniques like Monte Carlo analysis are required to account for statistical uncertainty in system parameters [7], [12], [13]. If the models are complex, generating solutions may be very time consuming. More importantly, it can be difficult to relate a particular result (e.g., a high level of crosstalk in one circuit) to specific characteristics of the system, making it challenging to solve problems that are identified.

Closed-form approximations for crosstalk can be particularly useful for analysis early in the design process [14]. Closed-form approximations allow rapid analysis of very complex systems to identify specific problem areas. Because these solutions are closed-form, the relationship between input parameters and the resulting crosstalk is well defined, allowing the user to easily identify those system characteristics that are causing a problem and to intelligently change the system to fix them. Closed-form approximations are often used hand-in-hand with more complicated numerical models since closed-form solutions can be used to quickly evaluate thousands of possible system interactions and identify those that merit further investigation. Detailed numerical analysis can then be conducted using only that information that is required to analyze the problem, allowing more efficient use of the modeling tools and a better understanding of the results.

Closed-form solutions for crosstalk based on transmission line theory can be found in the literature [15]–[20]. When looking for problem areas, however, one often wants to find the worst (or nearly worst) case, which is not reasonable with most published formulas. Worst case estimates are available for low frequencies, where circuits are electrically short [14], but typically not for high frequencies, where circuits are electrically long. Closed-form estimates of the worst case crosstalk at high frequencies are presented in [21], but these solutions are limited to a few specific configurations. Existing closed-form solutions at high frequencies also assume that the culprit and victim circuits are of the same length, which is often not the case in real harnesses.

In this paper, simple formulas are derived to approximate the worst case crosstalk between a culprit and victim circuit at high frequencies, where the wavelength is small compared to the length of the circuit. The paper begins by deriving an expression for the maximum crosstalk between two lossless transmission lines. The worst case crosstalk is calculated by assuming that energy is coupled at the "worst" location and excites resonances in the victim. Approximations are validated using bench-top experiments and using measurements taken in an automobile.

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II. Theory

Approximations for the worst case crosstalk between coupled transmission lines were made using the following assumptions:

1) The geometry is uniform where coupling occurs.
2) The medium is homogeneous.
3) The transmission lines are weakly coupled.
4) The transmission lines are lossless.

A uniform geometry assumption is required to avoid complex formulations as the geometry changes along its length. Assuming a homogeneous medium allows the use of simple closed-form formulas to predict coupling. While cable harnesses are rarely homogeneous, due to the differing permittivity of wire insulation and air, it is relatively common to replace the inhomogeneous medium with a homogenous medium whose characteristics approximate the inhomogeneous case. Assuming the transmission lines are weakly coupled allows one to ignore the coupling of noise from the victim back to the culprit. This is generally a safe assumption in most automotive wire harnesses; however, if the coupling is not weak, the closed-form expressions derived here can still be used to identify a coupling problem. The lossless transmission line assumption is reasonable if overall losses are dominated by losses at the terminations, which is the case in most practical configurations.

When circuits are electrically long, the voltage and current are not constant along the length of the circuit. In the worst case, the peak voltage or current in the culprit circuit will be in a position that most efficiently drives the victim circuit, the victim will be at an appropriate length to resonate, and the peak voltage or current will appear at the victim circuit load. This situation will not occur at all frequencies, but there is a reasonable likelihood that it will occur (or nearly occur) at least at some frequencies when the circuit is electrically long.

To simplify the following derivations, an alternative expression for crosstalk is defined here from the ratio of maximum voltage or current in the circuits as $X_{\mathrm{MAX}} = V_{2\mathrm{MAX}}/V_{1\mathrm{MAX}}$ or $X_{\mathrm{MAX}} = I_{2\mathrm{MAX}}/I_{1\mathrm{MAX}}$, where X_{MAX} is the worst case high-frequency crosstalk, $V_{2\mathrm{MAX}}$ and $I_{2\mathrm{MAX}}$ are the magnitude of the maximum voltage and current along the victim circuit, and $V_{1\mathrm{MAX}}$ and $I_{1\mathrm{MAX}}$ are the magnitude of the maximum voltage and current along the culprit circuit. This definition accounts for the fact that voltage and current may change along the harness length. Knowledge of the maximum voltage or current along the length of the wire can be used to estimate voltage at the culprit or victim loads when circuit parameters are known, allowing this definition to be linked back to more traditional definitions of crosstalk.

The worst case crosstalk is found by first determining how the maximum voltage or current along the culprit will couple to the victim circuit. This coupled noise is modeled as a lumped source that drives a transmission line representing the victim. The worst case crosstalk is then found by varying the position of the lumped noise source until it drives the maximum voltage or current along the length of the victim.

Consider two transmission lines of infinite length that are weakly coupled, so that the current and voltage in the culprit are minimally influenced by the induced current and voltage in the victim. In this case, the ratio of the magnetic flux wrapping the victim circuit to the magnetic flux wrapping the culprit circuit is given by l_m/l_{11} [20]. Without considering the terminating impedances of the victim circuit, the ratio of the maximum voltage along the length of the victim circuit to the maximum voltage along the length of the culprit circuit is $V_{2\mathrm{MAX}}/V_{1\mathrm{MAX}} \approx l_m l_{11}$. Similarly, the ratio of the maximum current on the two transmission lines is given by $I_{2\mathrm{MAX}}/I_{1\mathrm{MAX}} \approx C_m/(C_{22} + C_m)$. For transmission lines in a homogeneous medium, inductance and capacitance are directly related as $l_m/l_{11} = C_m/(C_{22} + C_m)$ [20], so that $V_{2\mathrm{MAX}}/V_{1\mathrm{MAX}} \approx I_{2\mathrm{MAX}}/I_{1\mathrm{MAX}}$. This relationship does not precisely hold for an inhomogeneous medium, for example, when wires are surrounded by polyvinyl chloride (PVC) insulation and suspended in air, but should be appropriate here for approximating maximum crosstalk.

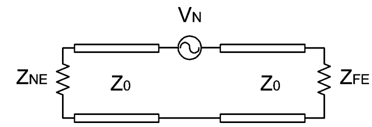

Fig. 1. Equivalent circuit model of weak inductive coupling to the victim circuit at a single location.

As mentioned earlier, the maximum crosstalk is assumed to be caused by coupling at an "ideal" location along the transmission line where there are large voltages or currents in the culprit that may easily drive the victim. The coupling at this arbitrary location can be represented as a "noise source" that drives the victim as shown in Fig. 1. The noise source can be represented as either a voltage source, associated with magnetic field coupling as $|V_N| \approx V_{1\mathrm{MAX}}(l_m/l_{11})$, or as a current source, associated with electric field coupling as $|I_N| \approx I_{1\mathrm{MAX}}(C_m/(C_{22} + C_m))$. Either source can be used since inductance and capacitance are directly related in a homogeneous transmission line. Here we estimate the maximum crosstalk using a voltage source.

The maximum coupled noise voltage in the victim circuit can be found by starting with the circuit in Fig. 1. The noise voltage is set to the maximum voltage coupled along an infinite length transmission line from the culprit, $|V_N| \approx V_{1\mathrm{MAX}}\,(l_m/l_{11})$. Z_0 is the characteristic impedance of the victim circuit, and Z_{NE} and Z_{FE} are the near-end and far-end termination impedances of the victim circuit. To calculate the maximum noise voltage that occurs along the finite-length victim circuit, the voltage to the right and left of the noise source must be derived. The maximum voltage to the right of the noise source is calculated as shown in Fig. 2, where Z_{LHS} is the impedance looking into the left-hand side (LHS) of the circuit from the noise source and x_0 is the length of the circuit from the noise source to the far end. Note that the noise source may be at an arbitrary location along the circuit's length.

The voltage and current along a lossless transmission line are given by [23]

$$V(x) = V^+(1 + \Gamma_{\mathrm{FE}} e^{j2\beta x})e^{-j\beta x} \tag{1}$$

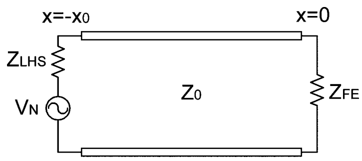

Fig. 2. Equivalent circuit used to calculate the maximum voltage on the victim circuit, to the right of the noise source.

and

$$I(x) = \frac{1}{Z_0} V^+ (1 - \Gamma_{\mathrm{FE}} e^{j2\beta x}) e^{-j\beta x} \tag{2}$$

where V^+ is the incident voltage wave amplitude (from source to load), x is position, $\Gamma_{\mathrm{FE}} = Z_{\mathrm{FE}} - Z_0 / Z_{\mathrm{FE}} + Z_0$ is the reflection coefficient at the far end of the circuit ($x = 0$), $Z_{\mathrm{FE}} = R_{\mathrm{FE}} + jX_{\mathrm{FE}}$ is the far-end load impedance of the victim circuit, and $\beta^2 = \omega^2 me$.

At the near end, where $x = -x_0$,

$$V(x = -x_0) = V_N - Z_{\mathrm{LHS}} I(x = -x_0). \tag{3}$$

Substituting (1) and (2) into (3),

$$V^+ (1 + \Gamma_{\mathrm{FE}} e^{-j2\beta x_0}) e^{j\beta x_0} = V_N - \frac{Z_{\mathrm{LHS}}}{Z_0} V^+ \times (1 - \Gamma_{\mathrm{FE}} e^{-j2\beta x_0}) e^{j\beta x_0}. \tag{4}$$

Rearranging and solving for V^+ yields

$$V^+ = \frac{V_N}{2} \left(1 - \frac{Z_{\mathrm{LHS}} - Z_0}{Z_{\mathrm{LHS}} + Z_0}\right) \times \frac{e^{-j\beta x_0}}{1 - (Z_{\mathrm{LHS}} - Z_0)/((Z_{\mathrm{LHS}} + Z_0))\Gamma_{\mathrm{FE}} e^{-j2\beta x_0}}. \tag{5}$$

Defining $\Gamma_{\mathrm{LHS}} = Z_{\mathrm{LHS}} - Z_0 / Z_{\mathrm{LHS}} + Z_0$, the aforementioned equation becomes

$$V^+ = \frac{V_N}{2} (1 - \Gamma_{\mathrm{LHS}}) \frac{e^{-j\beta x_0}}{1 - \Gamma_{\mathrm{LHS}} \Gamma_{\mathrm{FE}} e^{-j2\beta x_0}}. \tag{6}$$

However, we also have [23]

$$\Gamma_{\mathrm{LHS}} = \Gamma_{\mathrm{NE}} e^{-j2\beta(l - x_0)} \tag{7}$$

where $\Gamma_{\mathrm{NE}} = Z_{\mathrm{NE}} - Z_0 / Z_{\mathrm{NE}} + Z_0$ is the reflection coefficient at the near end of the circuit, $Z_{\mathrm{NE}} = R_{\mathrm{NE}} + jX_{\mathrm{NE}}$ is the near end terminating impedance of the victim circuit, and l is the length of the circuit.

Substituting (7) into (6) gives

$$V^+ = \frac{V_N}{2} (1 - \Gamma_{\mathrm{NE}} e^{-j2\beta(l - x_0)}) \frac{e^{-j\beta x_0}}{1 - \Gamma_{\mathrm{NE}} \Gamma_{\mathrm{FE}} e^{-j2\beta l}}. \tag{8}$$

Substituting (8) into (1), one gets the general expression for the voltage on the victim circuit,

$$V(x) = \frac{V_N}{2} (1 - \Gamma_{\mathrm{NE}} e^{-j2\beta(l - x_0)}) \frac{(1 + \Gamma_{\mathrm{FE}} e^{j2\beta x}) e^{-j\beta(x + x_0)}}{1 - \Gamma_{\mathrm{NE}} \Gamma_{\mathrm{FE}} e^{-j2\beta l}}. \tag{9}$$

The worst case voltage along the victim circuit is thus given by

$$|V_{\mathrm{MAX}}| \approx \frac{|V_N|}{2} \frac{(1 + |\Gamma_{\mathrm{NE}}|)(1 + |\Gamma_{\mathrm{FE}}|)}{1 - |\Gamma_{\mathrm{NE}}| \, |\Gamma_{\mathrm{FE}}|}. \tag{10}$$

Using $|V_N| \approx V_{1\mathrm{MAX}} \frac{l_m}{l_{11}}$ (or $|V_N| \approx V_{1\mathrm{MAX}} (C_m / (C_{22} + C_m))$),

$$V_{\mathrm{MAX}} \approx V_{1\mathrm{MAX}} \frac{l_m}{l_{11}} \frac{(1 + |\Gamma_{\mathrm{NE}}|)(1 + |\Gamma_{\mathrm{FE}}|)}{2(1 - |\Gamma_{\mathrm{NE}}| \, |\Gamma_{\mathrm{FE}}|)} \tag{11}$$

or

$$V_{\mathrm{MAX}} \approx V_{1\mathrm{MAX}} \frac{C_m}{C_m + C_{22}} \frac{(1 + |\Gamma_{\mathrm{NE}}|)(1 + |\Gamma_{\mathrm{FE}}|)}{2(1 - |\Gamma_{\mathrm{NE}}| \, |\Gamma_{\mathrm{FE}}|)}. \tag{12}$$

The worst case crosstalk is, therefore,

$$X_{\mathrm{MAX}}(\mathrm{dB}) \approx 20 \log_{10} \left| \frac{l_m}{l_{11}} \frac{(1 + |\Gamma_{\mathrm{NE}}|)(1 + |\Gamma_{\mathrm{FE}}|)}{2(1 - |\Gamma_{\mathrm{NE}}| \, |\Gamma_{\mathrm{FE}}|)} \right| \tag{13}$$

or

$$X_{\mathrm{MAX}}(\mathrm{dB}) \approx 20 \log_{10} \left| \frac{C_m}{C_m + C_{22}} \frac{(1 + |\Gamma_{\mathrm{NE}}|)(1 + |\Gamma_{\mathrm{FE}}|)}{2(1 - |\Gamma_{\mathrm{NE}}| \, |\Gamma_{\mathrm{FE}}|)} \right|. \tag{14}$$

Equations (13) and (14) show that there are primarily two factors determining the worst case values of crosstalk. One is the self and mutual inductance of the culprit and victim circuits, which is determined by circuit geometry and the characteristics of the intervening media. The other is the source and load terminations of the victim circuit, which determine the loss and the quality factor of the resonance in the victim circuit. The resonant property of the victim circuit plays an important role in determining the worst case crosstalk. The worst case crosstalk can be high even when the coupling between the culprit and victim circuits is very weak (e.g. $C_m \approx C_{22}$, $l_m \approx l_{11}$) if the victim circuit exhibits significant resonance (i.e., $|\Gamma_{\mathrm{NE}}|$ and $|\Gamma_{\mathrm{FE}}|$ are both close to 1).

III. Validation

A. Bench-top Validation

Experiments were performed to show the accuracy of (14) on a bench-top setup and in an automobile wire harness. The bench-top setup is shown in Fig. 3. Culprit and victim circuits were created by suspending two wires approximately 1.5 cm over an aluminum plate using Styrofoam. Both circuits used the plate as the current return path. The source end of the culprit circuit was connected to port 1 of a network analyzer and the far or near end of the victim circuit was connected to port 2 of the network analyzer. S_{21} was measured and used to calculate crosstalk, taking into account the value of the termination impedances. The radius of the wires was 0.4 mm. The distance between the circuits was approximately 5 cm. The wires were made long so that experiments could be performed at relatively low frequencies where termination parasitics could be neglected. The length of the victim circuit was about 7.5 m, which is equivalent to a quarter of a wavelength at 10 MHz. The length of the culprit circuit was varied from 2 m to 7.5 m. For most measurements the culprit circuit was 7.5 m long and placed

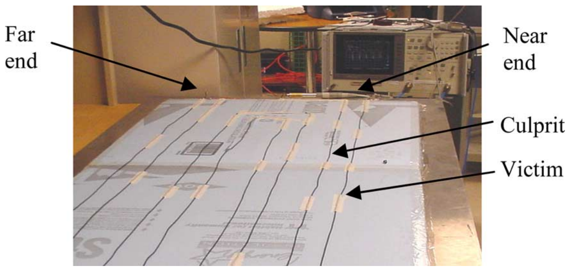

Fig. 3. Experimental setup.

TABLE I
TRANSMISSION LINE PARAMETERS

	C_{11}, C_{22} (pF/m)	C_m (pF/m)	l_{11}, l_{22} (μH/m)	l_m (μH/m)	Z_0 (Ω)
Measured	13	0.6	0.9	0.04	260
Calculated	13	0.5	0.9	0.03	260

TABLE II
CONFIGURATIONS OF CULPRIT AND VICTIM CIRCUITS

	Culprit		Victim		Estimation
	R_S	R_L	R_{NE}	R_{FE}	error
Case 1	250 Ω	10 kΩ (open)	0 Ω (short), 50 Ω, or 10 kΩ	10 kΩ (open)	≈ 6 dB
Case 2	250 Ω	10 kΩ (open)	250 Ω	1 kΩ or 10 kΩ (open)	≈ 2 dB
Case 3	250 Ω	0 Ω (short)	510 Ω, 1 kΩ, or 10 kΩ	0 Ω (short)	≈ 6 dB
Case 4	250 Ω	10 kΩ (open)	510 Ω, 1 kΩ, or 10 kΩ	0 Ω (short)	≈ 4 dB
Case 5	250 Ω	0 Ω (short)	0 Ω, 50 Ω, or 110 Ω	10 kΩ (open)	≈ 3 dB
Case 6	250 Ω	0 Ω (short)	240 Ω ∥ 47 pF	1 kΩ or 10 kΩ (open)	≈ 3 dB
Case 7	250 Ω	0 Ω (short) or 10 kΩ (open)	1 Ω (short)	1 Ω (short)	≈ 3 dB

beside the victim for its entire length. In some cases the culprit circuit was made much shorter than the victim and was placed at an arbitrary location along the victim's length. In these cases, the length and the position of the culprit circuit was varied to cause worst case crosstalk. The inductance and capacitance of the transmission lines were calculated and measured and are shown in Table I. Values for mutual inductance were calculated assuming that $l_m = C_m l_{11}/(C_{22} + C_m)$.

Seven circuit configurations were studied as shown in Table II. These configurations were chosen to test the performance of the approach in the extreme cases, rather than to closely match circuits typically found in a vehicle where crosstalk is typically lower. Experiments included several configurations where one or both ends of the circuit were nearly open or nearly short. Parasitic inductance or capacitance will similarly cause the terminations to have a very high or low impedance at high frequencies. Circuit lengths were also varied. The source impedance, R_S, of the culprit circuit was set to 250 Ω, which is approximately matched to the characteristic impedance of the transmission line. Nearly open circuits were achieved using a 1- or 10-kΩ resistor.

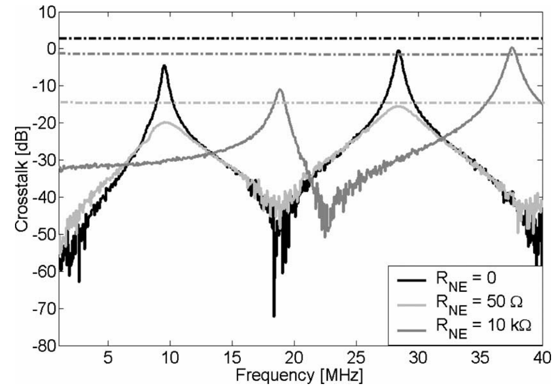

Fig. 4. Measured (solid line) and approximation of worst case crosstalk (dash-dotted line) when $R_L = 10$ kΩ (culprit) and $R_{\mathrm{FE}} = 10$ kΩ (victim).

In Case 7, nearly shorted circuits were achieved using a 1-Ω resistor. Due to the size of the terminations and the frequency, conductive and dielectric loss can be reasonably ignored for all circuits. Crosstalk was calculated from the measured values of S_{21}.

Figs. 3–5 show example values of measured crosstalk and the associated worst case estimates of crosstalk. In Fig. 4, the near-end termination of the victim was either $R_{\mathrm{NE}} = 0\Omega$, 50 Ω, or 10 kΩ, and the culprit circuit was about 2 m long starting at a point 2 m away from the near end of the victim circuit. Because neither end of the victim circuit was matched, the Q-factor of the resonance was relatively high. Fig. 5 shows results when the victim circuit was terminated with a capacitive load. Fig. 6 shows results when the victim circuit was terminated with low impedance at both ends. In this case, the maximum voltage will not occur across the source or load termination, but the maximum current does. For this reason, the current instead of the voltage was measured at the terminations for this low-impedance case and crosstalk was calculated as a ratio of the maximum current on the victim circuit to the maximum current on the culprit. The voltage across the 1-Ω resistor at the load end of the victim, and thus the current at the load end, was measured at port 2 of the network analyzer. Several other circuits were measured with similar results as shown in Table II and detailed in [24]. In all cases, the maximum values of crosstalk matched worst case predictions within about 6 dB or less.

B. Validation in an Automobile

Experiments were also performed on three circuits sharing the same harness bundle inside an automobile. Measurements were made on an existing harness bundle in the engine compartment, containing approximately 160 wires as shown in Fig. 7. To perform the test, the circuits were randomly chosen from the harness bundle. The length of the circuits was about 2 m. Each circuit tested consisted of two wires. Their relative placement in the harness was random and expected to change through the harness, though the circuits were neither particularly close nor far from one another at the near end. The return wires of the two circuits were shorted together at the near end and connected to the outer conductor of the SMA connectors as indicated in Fig. 8. The signal wires were connected to the inner conductor

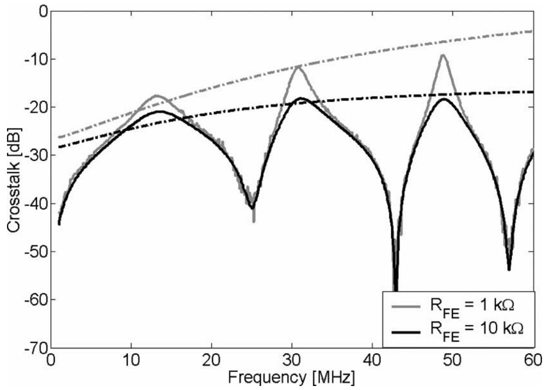

Fig. 5. Measured (solid line) and approximation of worst case crosstalk (dash-dotted line) when $R_L = 0$ (culprit), $R_{\rm FE} = 1$ kΩ or 10 kΩ (victim), and $Z_{\rm NE}$ = 240 Ω ∥ 47 pF (victim).

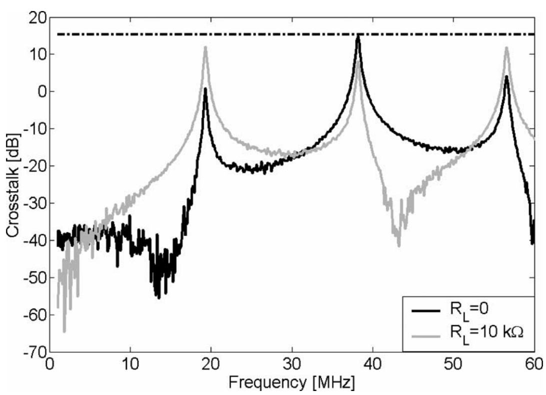

Fig. 6. Measured (solid line) and approximation of worst case crosstalk (dash-dotted line) when $R_L = 0$ or 10 kΩ (culprit) and $R_{\rm FE} = R_{\rm NE} = 1$ Ω (victim).

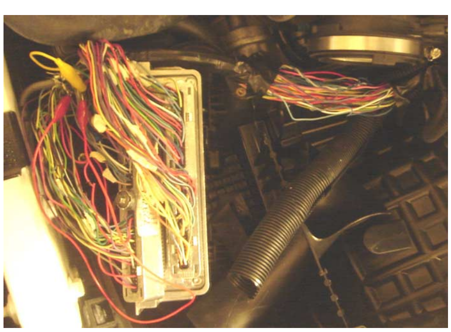

Fig. 7. Crosstalk was measured between circuits connected to the engine control module.

of the SMA connectors. The culprit circuit was named "circuit 1" and was terminated with a 100-Ω resistor at the source end and was either open or shorted at the load end. The victim circuits were named "circuit 2" and "circuit 3" and were shorted at the far end and terminated with a 10-kΩ resistor at the near end to test the extreme case. The parasitic parallel capacitance of the resistor was measured using a network analyzer and was about 0.33 pF. For these terminations, the transmission line can be approximated as lossless since the loss is dominated by the termination impedances. Since the position of the harness relative to body-surface metal was not known everywhere, transmission line characteristics were measured and the values obtained are shown in TABLE III. It is useful to note that the calculated values of mutual capacitance ranged from about 4 pF/m to 10 pF/m.

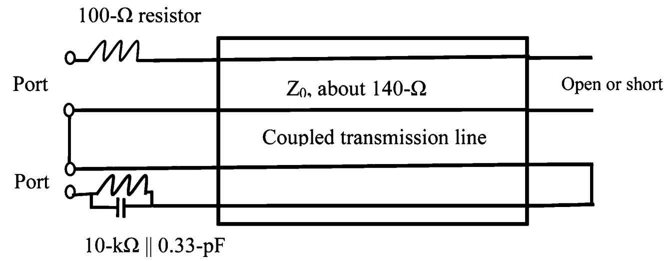

Fig. 8. Equivalent circuit of the measurement setup

TABLE III
MEASURED TRANSMISSION LINE PARAMETERS IN AUTOMOBILE

	Self capacitance (pF)	Mutual capacitance to circuit 1 (pF)	Self inductance (μH)	Characteristic impedance (Ω)
Circuit 1	92		1.9	140
Circuit 2	112	7.5	1.7	120
Circuit 3	110	5.5	1.6	120

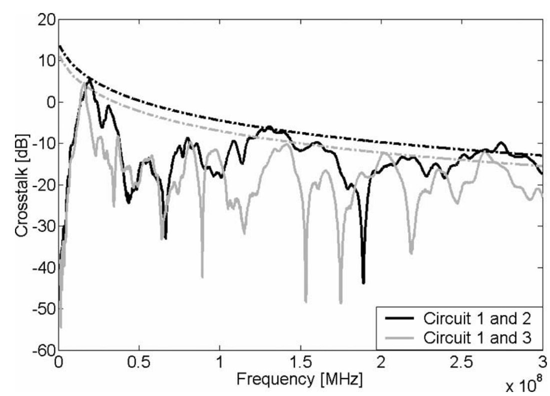

Fig. 9. Measured (solid line) and approximation of worst case intraharness crosstalk (dash-dotted line) inside an automobile, when the culprit circuit was terminated with an open at the load end.

Fig. 9 shows the measured and worst case estimates of crosstalk when the culprit circuit was open at the load end. Fig. 10 shows the results when the culprit circuit was shorted at the load end. In both cases, the maximum crosstalk was approximated to within about 3 dB.

IV. Discussion

The approximations derived here assume that the culprit circuit optimally excites resonances in the victim circuit. This may not occur in the actual circuit and maximum levels of crosstalk may be overestimated, depending on the circuit layout, circuit lengths, and other parameters. As these formulas assume resonance at every frequency, crosstalk at nonresonant frequencies will be overestimated. The influence of other circuits in a harness bundle is also ignored. Despite these limitations, the given approximations for worst case crosstalk are useful because they allow rapid determination of possible problem areas in the vehicle

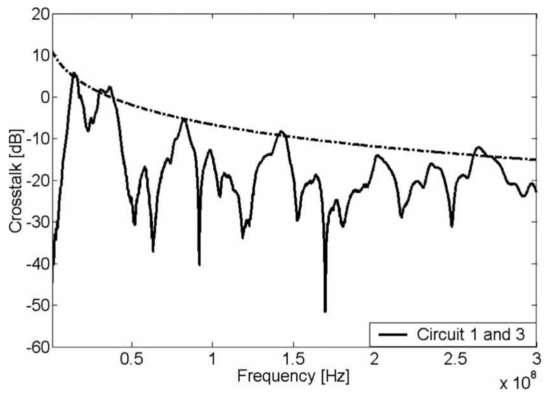

Fig. 10. Measured (solid line) and approximation of worst case intraharness crosstalk (dash-dotted line) inside an automobile, when the culprit circuit was terminated with a short at the load end.

using a simple, closed-form expression. As shown in the experimental results, the approximation is rough, but usually good enough to identify crosstalk issues. The equations are also simple enough that one can understand the most important parameters that influence crosstalk and intelligently modify problem circuits. The fact that the resonant frequencies are not given is not generally a problem, since the precise frequency of resonance may be difficult to determine due to the variations inherent in realistic harnesses, particularly in the presence of parasitics. If a more precise result is required for a particular culprit–victim pair, this pair can be analyzed using detailed numerical models.

The maximum crosstalk at high frequencies is determined by two factors, the strength of the coupling between the culprit and victim circuit, and the reflection coefficients in the victim, which is associated with resonance. The strength of the coupling can be reduced by separating the two circuits, by using shielding, or by other methods. A critical circuit should be carefully designed to avoid high-Q resonances, especially when it shares the same harness bundle with a circuit that may contain energy with spectral components that might excite this resonance. A matched source or load impedance in the victim circuit will significantly reduce the quality factor and thus improve the immunity of the circuit to crosstalk. The characteristics of circuits near a critical circuit should also be designed carefully. For example, signal return wires are sometimes connected to the automobile chassis at one or both ends. These "ground" wires can develop a high-Q resonance.

V. Conclusion

Simple formulas to approximate the worst case crosstalk between coupled transmission lines at high frequencies were derived and validated experimentally. While the formulas were created based on several simplifying assumptions, they proved to work reasonably well in realistic situations. For the scenarios studied here, maximum values of crosstalk were within 6 dB or less of each calculated or measured peak value. The formulas are especially well suited for crosstalk analysis early in the design process, where the exact value of input parameters may be unknown and rapid determination of potential crosstalk issues is particularly important.

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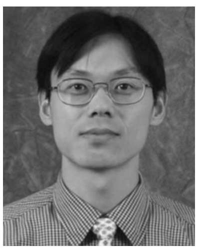

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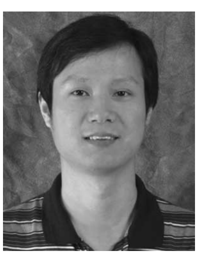

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