Passive Sensing Node Network for Monitoring Performance of RF Connector-Coaxial Cable System

1 Joseph F. REVELLI, 2 Noah MONTENA, and 1 Robert J. BOWMAN

1 ADIML Laboratory at Rochester Institute of Technology, Rochester, NY, 14623, USA
2 PPC, East Syracuse, NY, 13057, USA
Tel.: 585-475-7917, fax: 585-475-5845
E-mail: rjbeee@rit.edu

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Abstract: A passive sensing node technology is introduced for monitoring the integrity of RF coaxial transmission lines and associated RF connectors in the UHF range of frequencies. Each passive sensing node is powered by a transmission line coupler inside the coaxial cable. Passive sensing node sensing elements, fine-line interconnect metallurgy, transmission line coupler, and electronics are packaged on a molded interconnect device. The molded interconnect device structure and substrate material are designed to minimize electrical disturbance to the coaxial cable and preserve coaxial cable mechanical integrity. A custom integrated circuit on the molded interconnect device measures coaxial cable RF performance parameters, harvests operating power for the custom integrated circuit, and enables remote telemetry. The passive sensing node operates with a 60 µW power listening mode and a 900 µW power active mode for operating the three primary sensors, signal processing circuits, and backscatter telemetry. This paper proposes a network of passive sensing nodes in a coaxial cable system, describes the operation of the sensors in each passive sensing node, and provides a mathematical model for the transmission line coupler. Copyright © 2014 IFSA Publishing, S. L.

Keywords: Coaxial cable, RF connector, Radio frequency (RF), Passive sensing node (PSN), Molded interconnect device (MID).

1. Introduction

The RF (radio frequency) coaxial cable is a specialized controlled impedance medium used to transmit RF, video, high speed data, or precision measurement signals. The coaxial cable is configured with an insulated center conductor separated from a concentric shield providing an isolated and stable transmission medium. Sections of coaxial cable are connected together using RF connectors. RF connectors are designed to maintain the electrical shielding and characteristic impedance of the coaxial cable system. The telecommunications industry uses millions of RF connectors every year to interconnect sections of coaxial cables. The two primary failure modes in cellular tower RF connectors are connectors becoming loose and connector moisture ingress. Either condition can significantly alter the characteristic impedance of the RF transmission line and cause system degradation or catastrophic failure. In spite of considerable improvements in connector design, RF connector failures continue to plague the telecommunications industry.

A passive sensing node (PSN) is described to detect the dominant failure mechanisms in RF coaxial connectors. In addition, the PSN monitors the quality...
of propagating RF signals in the coaxial cable, harvests RF energy from the coaxial cable, and provides a telemetry technology to report both the system status and unique identification of each connector PSN. Degrading quality in an RF connector can be detected and corrected before catastrophic system failure occurs. The paper is organized to discuss the major subsystems of the PSN.

A typical coaxial cable system is depicted as a distributed passive sensor network in Section 2. The Molded Interconnect Device (MID), described in Section 3, is designed to package the passive sensor node. The operation of the primary sensors used for monitoring cable and connector integrity is discussed in Section 4. Section 5 analyzes the electromagnetic coupling characteristics of the intra-coaxial cable Transmission Line Coupler (TXC), and Section 6 summarizes and concludes.

2. The Distributed Sensor Network

Fig. 1 depicts a simplified view of a cable configuration for a cell tower coaxial cable installation. The antenna symbol is representative of a multiple antenna array. The radio transmitters and receivers in the equipment shed are connected to the antenna array through a network of coaxial jumpers and the main coaxial cable trunk. The main cable trunk traverses up the tower shaft over a distance of typically 50-200 feet depending on the tower height. The PSNs are located in each female side of the mated RF connectors along the coaxial cable, constituting a wireless, passive sensor network. Additional RF connector pairs (and therefore PSNs) are often installed on the coaxial cable to accommodate specialized equipment such as surge suppressors. A master unit in the equipment shed provides a dedicated RF signal for power harvesting and polling the individual PSNs for sensor status.

Cell towers and associated coaxial cable systems are shared resources. A single tower can accommodate many coaxial cable systems. In addition, multiple RF transmit and receive channels are frequency division multiplexed onto a single coax cable. The electrical behavior of the PSNs and associated master unit, which share the co-axial cable bandwidth, are designed not to interfere with or degrade the normal operation of the RF cellular channels.

3. Molded Interconnect Device Design

The significant advancements in the properties of plateable thermoplastics and the ability to deposit and pattern surface metallurgy on these plastics have made MIDs the preferred packaging technology in many micro-electromechanical systems [1-7]. The passive sensing node in this paper is packaged as a MID using a liquid crystal polymer (LCP) as the substrate material [8]. The webbed, disc-shaped MID shown in Fig. 2 is created with an injection molding process. The fine line metallurgy for electronic circuit interconnect and antenna definition is patterned with laser direct structuring [9].

Several substrate materials were tested and the LCP provided the desired mechanical properties to meet the compression, thermal, shock, metal surface adhesion, metal patterning, and vibration requirements. In addition, the LCP satisfies requirements for electrical properties including low dielectric constant and low leakage. The geometric structure of the MID was designed to provide a metal annular ring for sensing connector tightness, a reservoir cavity to contain the hydrophilic polymer Nafion® for moisture sensing, and a recessed landing zone on the outer ring for surface mounting the custom integrated circuit (IC). Placing the IC in an RF shielded pocket on the disc rim minimizes RF interference from the center conductor of the coax cable. The MID is designed to provide a 20 mil gap.
between the center conductor of the coaxial cable and the MID inner surface to prevent electric field discharge between the center conductor and the circuitry on the MID. A quarter cross-section view in Fig. 3 depicts the PSN MID mounted inside a 7/16 DIN connector.

The MID is press-fit into the connector housing at the time of manufacture. The outer surface of the MID disc is also metalized to establish an electrical connection to the cable shield [10]. The center conductor of the coaxial cable passes through the center hole of the MID but does not make direct electrical contact with the MID. Additional information on the structural design of the MID and the packaging of the integrated circuit including RF shielding can found in [8].

4. PSN Primary Sensors

The PSN acts a remote data acquisition unit and contains four electrical subsystems as shown Fig. 4.

4.1. RF Connector Tightness

RF connector tightness is determined using a capacitive Wheatstone bridge as shown in Fig. 5. All elements shown except $C_{\text{sense}}$ are included on the custom CMOS integrated circuit. $C_{\text{sense}}$ is an external capacitor formed by the annular ring on the MID and the metal surface on the connector that moves with the tightening of the mating RF connector [13]. $C_{\text{sense}}$ (which includes some parasitic capacitance) varies over the range of 5-30 pF as the RF connector is adjusted from loose to fully tight. A 20 kHz 3 Vpp sinusoidal signal stimulates the bridge. A differential amplifier senses and amplifies the offset voltage developed across interior nodes of the bridge in response to variations in $C_{\text{sense}}$.

4.2. Moisture Ingress

Moisture ingress is monitored by measuring the relative humidity (RH) of the RF connector chamber. Many techniques have been developed for small sensor hygrometry [14], the measurement of RH. The RH sensor in the PSN is a variable resistor placed outside the integrated circuit. This variable resistor is configured in a Wheatstone bridge circuit topology similar to Fig. 5 with C1-C3 replaced by R1-R3, each with a value of 100 K$\Omega$. The RH sensing resistor is fabricated on an MID surface recess using an interdigitated metal finger array coated with a Nafion hydrophilic polymer film.

The film conductivity varies with relative humidity and thus induces a change in the resistance of the interdigitated electrode which is inverse to relative humidity. Once again, an offset voltage proportional to the resistance bridge imbalance, and hence RH, is amplified by a differential amplifier. The sensor resistance response is linear over an RH range of 15%-85% as illustrated in Fig. 6. The Nafion polymer is known to be chemically inactive,

![Fig. 3. Quarter cross-section of a 7/16 DIN female connector shown with the MID press-fit into the body.](image)

![Fig. 4. PSN functional block diagram.](image)

![Fig. 5. Capacitive bridge for sensing RF connector tightness.](image)
similar to Teflon®, and therefore is very stable. Although high accuracy is not essential to sense a failure mode of high moisture, the RH profile data is useful when characterizing the RF connector environment over time.

![Graph showing sense resistance versus relative humidity for the Nafion sensor.](image)

**Fig. 6.** Sense resistance versus relative humidity for the Nafion sensor.

4.3. Temperature

On-chip temperature sensing is provided to allow for temperature compensation of transducing elements and to monitor the temperature environment of the RF connector body. A fixed bias current develops a forward bias voltage across a silicon p-n junction. The junction voltage exhibits fractional temperature coefficient of approximately -2 mV/°C.

5. Electromagnetic Interaction between TXC and Coaxial Cable Transmission Line

RF energy harvesting, RF power monitoring, and backscatter telemetry all depend on the coupling of energy between the near fields of the center conductor in the coaxial cable and the TXC on the MID. The interaction between the forward and reverse traveling waves precludes the use of simple lumped element circuit models for such structures [15]. Although an analysis based entirely on traveling waves and distributed elements yields exact results, these solutions do not yield compact expressions that would be useful for device design and other engineering applications. These analyses involve solving Maxwell’s equations using numerical simulators such as ANSOFT’s HFSS™ software. In order to address this problem, a four-port model was developed that combines features of both distributed and lumped elements. This model is based on coupled transmission line (CTL) theory and is capable of predicting, with reasonable accuracy over UHF frequencies, the coupling characteristics of the loop antenna and center conductor of the coaxial cable as well as the influence of the coupled antenna on the effective impedance of the main transmission line. This approach is similar to the one used in reference [16] to analyze the microwave characteristics of a singly split double ring resonator.

Fig. 7 (a) shows an idealized representation of the MID inserted inside a segment of coaxial cable and defines the port designations used in the CTL model (*i.e.*, P1-P4). Fig. 7 (b) shows a detail view of the TXC antenna structure in close proximity to the center conductor of the coaxial cable.

![Diagram of idealized intra-coaxial transmission line coupler showing: (a) transmission line cable segment with MID, (b) detail of TXC structure.](image)

This section is divided into three sub-sections. The basic four-port model is described in Section 5.1. In this section, expressions for Z and S matrices are derived in terms of frequency and fifteen model parameters. Five of these model parameters are can be computed directly from first principles. The remaining ten parameters are obtained by fitting the spectral S-parameter characteristics generated by the model to results obtained from HFSS. The fitting process for the basic model is presented in Section 5.2. Section 5.3 discusses the results of the fitting process.
5.1. Basic Coupled Transmission Line Model

The Basic CTL model makes the following five assumptions:
1. RF frequencies are in the range of 100 MHz to 4 GHz range where all dimensions of the TXC are much smaller than a wavelength;
2. Exchange of energy between the coax center conductor and the loop antenna is approximated by the interaction between a pair of coupled transmission line segments;
3. The return path for all transmission line segments is the coaxial cable shield;
4. All metals are perfect conductors and all transmission line segments are lossless, and
5. The TXC and the coaxial cable segment are symmetric about a plane that is perpendicular to the direction of propagation and passes through the center of the TXC.

The scope of the CTL model is bounded by the first three assumptions. The last two assumptions reflect the symmetry of the physical system and reduce model complexity.

Fig. 8 shows a schematic diagram of the symmetric four-port coupled transmission line model. Transmission is assumed to be along the positive z-axis and the TXC is indicated by the elements enclosed within the dashed lines.

The portion of the main coaxial cable transmission line that is included in the TXC is designated as Line 1. The segment of the square loop antenna that is closest and parallel to the center conductor of Line 1 is designated as Line 2. Lines 3 and 4 are assumed to be identical and represent transmission line segments that are parallel to the y-axis. These lines connect the square loop antenna to Ports 3 and 4, respectively. Lines 1-4 are characterized by distributed inductances $L_{01}$, $L_{02}$, $L_{03}$ and distributed capacitances $C_{01}$, $C_{02}$, $C_{03}$, respectively, as indicated in the figure. Ports 1 and 2 are connected to either side of Line 1 via segments of unperturbed transmission line coaxial cable. These transmission lines are characterized by impedance $Z_{01}$ and phase delay $\varepsilon_{01}$. Ports 3 and 4 are connected to Lines 3 and 4, respectively, via segments of a different transmission line which is characterized by impedance $Z_{03}$ and phase delay $\varepsilon_{03}$.

The quantities $a_k$ and $b_k$ ($k=1, 2, 3, 4$) represent amplitudes of electromagnetic waves traveling into and out of, respectively, Ports 1-4. Lines 1-2 and Lines 3-4 represent two sets of coupled transmission lines. Coupling for the two sets of lines occurs over the lengths $l_1$ and $l_3$ with distributed mutual coupling inductances $L_{m01}$ and $L_{m03}$ and distributed mutual coupling capacitances $C_{m01}$ and $C_{m03}$, respectively. The capacitance, $C_T$, is a lumped-element tuning capacitor.

Coupled transmission Lines 1 and 2 are redrawn in Fig. 9 (a). Currents $I_1(z)$ and $I_2(z)$ flow in Lines 1 and 2, respectively, and voltages $V_1(z)$ and $V_2(z)$ appear on the respective lines. A time dependency of exp(\text{i}wt) has been assumed. The origin of the z-axis is defined so that the currents flowing into nodes $A_1$ and $A_2$ are $I_1(0)$ and $I_2(0)$ and the voltages at these nodes are $V_1(0)$ and $V_2(0)$, respectively. Likewise, the currents flowing out of nodes $B_1$ and $B_2$ are $I_1(l_1)$ and $I_2(l_1)$, respectively, and the node voltages are $V_1(l_1)$ and $V_2(l_1)$.

![Fig. 8. Symmetric four-port transmission line model for the TXC.](image)

![Fig. 9: Diagrams for modeling coupled transmission line segments: a) Lines 1 and 2, and b) Lines 3 and 4.](image)
Coupled transmission line theory [17] describes the propagation of electromagnetic waves across a region where two transmission lines are coupled. The voltages and currents in Lines 1 and 2 are related by the following four equations:

\[ \frac{\partial V_1}{\partial z} = -j\omega L_{01} I_1 - j\omega M_{01} I_2 \]  
\[ \frac{\partial V_2}{\partial z} = -j\omega L_{02} I_1 - j\omega M_{02} I_2 \]  
\[ \frac{\partial I_1}{\partial z} = -j\omega C_{01} V_1 + j\omega M_{01} V_2 \]  
\[ \frac{\partial I_2}{\partial z} = +j\omega C_{02} V_1 - j\omega C_{02} V_2 \]  

Since the coupling length \( l_1 \) has been assumed to be much smaller than the wavelength of the propagating wave, multiplying both sides of Equations 1-4 by \(-\Delta z\approx -l_1\) results in the following four approximations:

\[ \tilde{I}_k \approx \frac{1}{2}[I_k(0)+I_k(l_1)] \]  
\[ \tilde{V}_k \approx \frac{1}{2}[V_k(0)+V_k(l_1)] \]

where \( s = j\omega \) and \( k = 1, 2 \) represent average values of \( I_k(z) \) and \( V_k(z) \) over the interval \( \Delta z \).

Another four equations can be obtained for coupled Lines 3 and 4 shown in Fig. 9(b) by applying similar reasoning. In order to simplify notation, currents and voltages are defined with respect to the eight nodes shown in Fig. 9(a) and Fig. 9(b). The currents \( I_{A1}, \) etc., are defined as currents flowing into node \( A_1, \) etc., and voltages \( V_{A1}, \) etc., are the voltages at the nodes. The following eight equations are obtained using this notation:

\[ V_{A1} - V_{B1} = sL_{11}(I_{A1} - I_{B1})/2 + sL_{12}(I_{A2} - I_{B2})/2 \]  
\[ V_{A2} - V_{B2} = sL_{21}(I_{A1} - I_{B1})/2 + sL_{22}(I_{A2} - I_{B2})/2 \]

Two additional equations can be obtained by applying Kirchoff’s voltage and current laws to nodes \( C \) and \( D \) shown in Fig. 9(a):

\[ V_{C1} = V_{D1} \]  
\[ V_{C2} = V_{D2} \]

Furthermore, since nodes \( A_2, \) \( C \) and \( E_3 \) are all at the same potential, as are nodes \( B_2, \) \( D \) and \( E_4, \) it can be seen that

\[ V_{A2} = V_{E3} \]  
\[ V_{B2} = V_{E4} \]

As indicated in Fig. 8, the TXC includes all elements inside the dashed lines. Accordingly, the \( Z \)-matrix, \( Z_{\text{TXC}} \), of the TXC relates the four currents \( I_{A1}, I_{B1}, I_{E3}, \) and \( I_{F4} \) to the four voltages \( V_{A1}, V_{B1}, V_{E3}, \) and \( V_{F4} \) by the matrix equation.
There are a total of sixteen variables (eight currents and eight voltages) in Equations 9-20. After a considerable amount of algebra, eight of these twelve equations can be used to eliminate all variables except the eight appearing in Equation 21 and the remaining four equations can be manipulated to yield expressions for the matrix elements of \( Z_{TXC} \).

It should be noted here that some of the complexity associated with the algebraic manipulations can be alleviated by taking advantage of the assumptions mentioned previously. In particular, it can be seen immediately that the assumed symmetry of the model results in

\[
Z_{11} = Z_{22}, \quad Z_{33} = Z_{44}, \quad Z_{12} = Z_{21}, \quad Z_{34} = Z_{43}.
\]

Furthermore it can be seen that \( Z_{13} = Z_{13} \) and \( Z_{14} = Z_{14} \) by virtue of the reciprocity theorem. Consequently only six elements of the \( Z \)-matrix remain unique, which are arbitrarily designated as \( Z_{11}, Z_{12}, Z_{13}, Z_{14}, Z_{33}, \) and \( Z_{34} \). Likewise, the \( S \)-matrix also exhibits six unique elements: \( S_{11}, S_{12}, S_{13}, S_{14}, S_{33}, \) and \( S_{34} \).

The \( S \)-matrix for the TXC can be obtained from \( Z_{TXC} \) by the usual transformation

\[
S_{TXC} = \left[ \begin{array}{cccc} Z_{01} & 0 & 0 & 0 \\ 0 & Z_{01} & 0 & 0 \\ 0 & 0 & Z_{03} & 0 \\ 0 & 0 & 0 & Z_{03} \end{array} \right]^{1/2} \left[ \begin{array}{cccc} Z_{01} & 0 & 0 & 0 \\ 0 & Z_{01} & 0 & 0 \\ 0 & 0 & Z_{03} & 0 \\ 0 & 0 & 0 & Z_{03} \end{array} \right]^{1/2} - 1 \right].
\] (22)

where

\[
S_{TXC} = \left[ \begin{array}{cccc} Z_{01} & 0 & 0 & 0 \\ 0 & Z_{01} & 0 & 0 \\ 0 & 0 & Z_{03} & 0 \\ 0 & 0 & 0 & Z_{03} \end{array} \right]^{1/2} \left[ \begin{array}{cccc} Z_{01} & 0 & 0 & 0 \\ 0 & Z_{01} & 0 & 0 \\ 0 & 0 & Z_{03} & 0 \\ 0 & 0 & 0 & Z_{03} \end{array} \right]^{1/2} - 1 \right].
\] (23)

is the characteristic impedance matrix for the four transmission lines feeding the TXC, and \( I \) is the \( 4 \times 4 \) identity matrix. It can be shown that the \( S \)-matrix measured from the external ports 1-4 (see Fig. 8) can be obtained from \( S_{TXC} \) by the transformation

\[
S_{EXT} = M_o S_{TXC} M_o
\] (24)

where

\[
M_o = \left[ \begin{array}{cccc} e^{-j\omega t_0} & 0 & 0 & 0 \\ 0 & e^{-j\omega t_0} & 0 & 0 \\ 0 & 0 & e^{-j\omega t_3} & 0 \\ 0 & 0 & 0 & e^{-j\omega t_3} \end{array} \right]
\] (25)

is the phase delay matrix for the four transmission lines feeding the TXC.

Following the procedures outlined above, the six independent elements of \( Z_{TXC} \) and \( S_{TXC} \) can be expressed in terms of the complex frequency, \( s = j\omega \), and fifteen model parameters. These expressions turn out to be much too cumbersome to be presented here. However, there are several rather interesting features that should be mentioned. For example, each element of \( Z_{TXC} \) can be expressed as the product of \( s^4 \) times a ratio of two even-powered polynomials in \( s \).

\[
Z_{jk} = \frac{1}{s} \frac{N_{jk}(s)}{D(s)}.
\] (26)

The polynomial \( D(s) \) is common to all elements of \( Z_{TXC} \) and is of order 6 as are \( N_{13}(s), N_{14}(s), N_{33}(s) \) and \( N_{34}(s) \). The remaining two elements, \( N_{11}(s) \) and \( N_{12}(s) \), are 8th order polynomials. It should be noted that when finite conductivities are introduced in the model for the metallization of the TXC, both odd and even powers of \( s \) appear in the expressions for \( D(s) \) and \( N_{1d}(s) \).

The quantity \( Z_{34}^{TXC}(s) \) is defined as the impedance between nodes \( F_3 \) and \( F_4 \) and the quantity \( Z_{34}^{EXT}(s) \) is defined as the impedance between the center conductors of ports 3 and 4, respectively, (see Fig. 9(b)). In both definitions Ports 1 and 2 are terminated by the characteristic impedance \( Z_{01} \).

\[
Z_{34}^{TXC}(s) = \frac{V_{F3} - V_{F4}}{I_{F3}} = 2 \left[ \frac{Z_{33}(s) - Z_{34}(s) - (Z_{13}(s) - Z_{14}(s))^2}{Z_{01} + Z_{11}(s) - Z_{12}(s)} \right].
\] (27)

It can be seen immediately from Equations 26 and 27 that \( Z_{34}^{TXC}(s) \) is proportional to \( 1/D(s) \) so that roots of \( D(s) \) correspond to poles of \( Z_{34}^{TXC}(s) \). For typical values of the fifteen model parameters, four of the six roots of \( D(s) \) turn out to be purely imaginary (i.e., two sets of conjugates) and two are purely real. Furthermore, only one set of conjugate roots corresponds to a frequency in the range of interest (i.e. 100 MHz to 5 GHz). The expression for the circular frequency associated with this root is given by

\[
\omega_{14}^{TXC} = \sqrt{\frac{1}{4} + \frac{Z_{14}}{Z_{01}}} \sqrt{\left( \frac{Z_{14} - \frac{1}{4}}{Z_{01}} \right)^2 + \frac{Z_{13}^2}{4} + \eta \left( \frac{1}{4} \right)^2}
\] (28)

where
\[ \omega_{o} = \frac{1}{\sqrt{L_{2}C_{T}}} \]  
(28a)

\[ \omega_{3} = \frac{1}{\sqrt{(L_{3} - L_{m})(C_{3} + C_{m})}} \]  
(28b)

\[ \chi = \frac{\omega_{3}^{2}}{\omega_{o}^{2}} \]  
(28c)

\[ \eta = (C_{1} + C_{m})/C_{T} \]  
(28d)

It turns out that the quantity under the outside radical in Equation 28 is approximately equal to one so that \(\omega_{3} \approx \omega_{o}\). Accordingly, for frequencies near \(\omega_{3}^{4}\), \(Z_{34}^{TXC}(j\omega)\) can be approximated by

\[ Z_{34}^{TXC}(j\omega) \approx \frac{K}{\left(\omega_{NF}^{4}\right)^{2} - \omega^{2}} \approx \frac{K}{\omega_{o}^{2} - \omega^{2}} \]  
(29)

where \(K\) is the approximately constant for frequencies near the resonance. Equation (29) will be recognized as the impedance of an LC tank circuit where \(L=L_{2}\) and \(C=C_{T}\). \(Z_{34}^{EXT}(s)\) exhibits a similar resonance at \(\omega = \omega_{EXT}^{4}\) which differs slightly from \(\omega_{3}^{4}\).

5.2. Parameter Extraction

The Ansoft High Frequency Structure Simulator (HFSS™) is a high-performance full-wave electromagnetic field solver that employs finite element analysis to simulate electrical characteristics of arbitrary 3D volumetric passive devices. HFSS was used to simulate two different idealized intra-coaxial antenna couplers. In TXC-1 the square loop antenna structure is suspended inside a segment of coaxial cable transmission line with an air dielectric. TXC-2 differs from TXC-1 in that the square loop antenna is supported by a dielectric substrate. TXC-2 is illustrated in Fig. 7(a). The coaxial cable segment is terminated at ports P1 and P2 and RF energy is coupled to the outside at ports P3 and P4. Dimensions for the various geometric parameters defined in Fig. 7(a) and Fig. 7(b) are given in Table 1. The relative electric permittivity of the substrate material in TXC-2 was assumed to be 3.0.

Table 1. Dimensions of geometric parameters used in HFSS simulation (cm).

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>(r_{01})</td>
<td>0.4800</td>
</tr>
<tr>
<td>(r_{03})</td>
<td>0.0030</td>
</tr>
<tr>
<td>(R_{03})</td>
<td>1.1051</td>
</tr>
<tr>
<td>(R_{03})</td>
<td>0.0104</td>
</tr>
<tr>
<td>(l_{L})</td>
<td>2.3984</td>
</tr>
<tr>
<td>(l_{L3})</td>
<td>0.2754</td>
</tr>
<tr>
<td>(l_{1})</td>
<td>0.2032</td>
</tr>
<tr>
<td>(l_{2})</td>
<td>0.1000</td>
</tr>
<tr>
<td>(d)</td>
<td>0.0780</td>
</tr>
<tr>
<td>(t)</td>
<td>0.0034</td>
</tr>
</tbody>
</table>

Coefficients that appear in Equations 9-16 include the fifteen parameters \(L_{1}, L_{2}, L_{3}, L_{m}, C_{1}, C_{2}, C_{3}, C_{w}, Z_{01}, \xi_{01}, Z_{03}, \xi_{03}\) and \(C_{T}\). The lumped-element tuning capacitor, \(C_{T}\), is assumed to be given and the four coaxial cable parameters \(Z_{01}, \xi_{01}, Z_{03}\) and \(\xi_{03}\) can be computed from the physical characteristics of the unperturbed cable segments. For example, for cylindrical coaxial transmission lines \(Z_{01}\) and \(\xi_{01}\) can be computed from the well-known relationships

\[ Z_{01} = \frac{1}{2\pi} \sqrt{\frac{\mu_{0}}{\varepsilon_{01}}} \ln \frac{R_{01}}{r_{01}} \]  
(30)

\[ \xi_{01} = \frac{1}{2\pi} \sqrt{\mu_{0}\varepsilon_{01}} \]  
(31)

where \(\mu_{0}\) and \(\varepsilon_{01}\) are, respectively, the magnetic permeability of free space and the electric permittivity of the material filling the coaxial cable. \(Z_{03}\) and \(\xi_{03}\) can be obtained in a similar fashion. Using the dimensions in Table 1 and assuming air-filled couplers, the values \(Z_{01} = 50\ \Omega, \xi_{01} = 8.00\times10^{-11}\ \text{sec}, Z_{03} = 75\ \Omega, \xi_{03} = 3.34\times10^{-12}\ \text{sec} were obtained.

The remaining ten parameters were obtained by fitting model results to the results from HFSS simulations using a least-means-square algorithm. The results of parameter extraction are given in Table 2. These fits were obtained for three values of tuning capacitor: \(C_{T} = 5, 10, \text{ and } 15\ \text{pF}\). Note the entry \(-0\) for the value of \(C_{2}\) for TXC-1. The algorithm initially yielded a small negative value for \(C_{2}\). In order to maintain physically realizable parameter values, the algorithm was run again with \(C_{2}\) set equal to zero.

Table 2. Extracted model parameters.

<table>
<thead>
<tr>
<th>Inductances (nH)</th>
<th>(L_{1})</th>
<th>(L_{2})</th>
<th>(L_{3})</th>
<th>(L_{m})</th>
<th>(L_{m3})</th>
</tr>
</thead>
<tbody>
<tr>
<td>TXC-1</td>
<td>0.334</td>
<td>5.672</td>
<td>5.894</td>
<td>0.121</td>
<td>4.650</td>
</tr>
<tr>
<td>TXC-2</td>
<td>1.055</td>
<td>5.673</td>
<td>4.911</td>
<td>0.119</td>
<td>3.481</td>
</tr>
<tr>
<td>Capacitances (fF)</td>
<td>(C_{1})</td>
<td>(C_{2})</td>
<td>(C_{3})</td>
<td>(C_{m})</td>
<td>(C_{m3})</td>
</tr>
<tr>
<td>TXC-1</td>
<td>175.5</td>
<td>-0</td>
<td>123.0</td>
<td>63.6</td>
<td>42.1</td>
</tr>
<tr>
<td>TXC-2</td>
<td>660.3</td>
<td>131.4</td>
<td>383.3</td>
<td>116.5</td>
<td>308.0</td>
</tr>
</tbody>
</table>

Fig. 10 (a)-(f) show plots of the absolute values of the six independent \(S\)-parameters as functions of frequency over the range 0-4 GHz. Each figure compares results of the CTL model with corresponding results obtained from the HFSS simulations. A tuning capacitor of 10 pF for both the air-filled coupler (TXC-1) and the coupler with the dielectric substrate (TXC-2) was assumed for all plots. It can be seen that the agreement is quite good in all cases except perhaps for frequencies approaching 4 GHz. Fig. 11(a) and Fig. 11(b) compare results from the model and from HFSS for
TXC-2 for three different tuning capacitors: \( C_T = 5 \) pF, \( C_T = 10 \) pF, and \( C_T = 15 \) pF. These figures show \(|S_{13}|\) and \(|S_{33}|\) over the frequency range 0-2 GHz.

Fig. 10. S-parameter spectra for \( C_T = 10 \) pF. Solid lines are obtained with the CTL model and dots with HFSS simulation for TXC-1; dashed lines are obtained with the CTL model and crosses with HFSS simulation for TXC-2: (a) \(|S_{11}|\), (b) \(|S_{12}|\), (c) \(|S_{13}|\), (d) \(|S_{14}|\), (e) \(|S_{33}|\), and (f) \(|S_{34}|\).

Fig. 11. Comparison of (a): \(|S_{13}|\) and (b): \(|S_{33}|\) for TXC-2, for frequencies below 2 GHz, and for three values of \( C_T \); \( C_T = 5 \) pF (solid lines for CTL model, dots for HFSS simulation), \( C_T = 10 \) pF (dashed lines for CTL model, crosses for HFSS simulation), and \( C_T = 15 \) pF (dash-dotted lines for CTL model, squares for HFSS simulation).

### 5.3. Results of the Fitting Process

Each of the ten extracted model parameters has a different effect on the features of the electrical characteristics of the intra-coaxial antenna coupler. For example, as discussed previously, \( L_2 \) plays a dominant role in determining the resonance frequencies \( \omega_{14}^\text{TXC} \) and \( \omega_{14}^\text{EXT} \). Indeed, \( L_2 \) controls the resonance frequencies of all S-parameters. \( L_m \), on the other hand, affects only the amplitudes of these resonances. \( L_1 \), \( C_T \) and \( C_m \) seem to affect the slopes of
\[ |S_{11}(s)| \text{ and } |S_{21}(s)| \text{ while } C_{m} \text{ primarily controls the slopes of } |S_{12}(s)| \text{ and } |S_{14}(s)|. \] The parameters \( L_3, L_{m3}, C_3, C_{m3}, \) and \( C_2 \) all affect the frequency at which \( |S_{13}(s)| \) exhibits a zero (see Fig. 10(c)). In fact \( |S_{13}(s)| \) has no zero if coupled transmission Line 3 and Line 4 are not included in the model. \( C_2 \) also affects the high frequency behaviors of \( |S_{12}(s)|, |S_{13}(s)|, \) and \( |S_{14}(s)| \).

The inductance and capacitance of a segment of unperturbed air-filled coaxial cable of length \( l_1 \) can be computed from the well-known equations

\[ L_1 = \frac{\mu_0 l_1}{2\pi} \ln \frac{R_{01}}{r_{01}} \approx 0.333 \text{ nH} \quad (32) \]

and

\[ C_1 = \frac{2\pi \varepsilon_0}{\ln \frac{R_{01}}{r_{01}}} \approx 135.6 \text{ fF} \quad (33) \]

for values of \( l_1, R_{01} \) and \( r_{01} \) listed in Table 1. Although the computed inductance is very close to the fit value of \( L_1 = 0.334 \text{ nH} \) given in Table 2 for TXC-1, the computed capacitance is smaller than the fit value of \( C_1 = 175.5 \text{ fF} \) obtained for TXC-1. It is also interesting to note that while the fit values of \( L_1 \) and \( C_1 \) are dramatically different for TXC-1 and TXC-2, the characteristic impedance obtained from the fitting process, \( Z_{1,\text{fit}} = |Z_1|/C_1 \), is nearly the same for the two devices: \( Z_{1,\text{fit}} \approx 43.6 \Omega \) for TXC-1 and \( Z_{1,\text{fit}} \approx 40 \Omega \) for TXC-2.

#### 6. Summary and Conclusion

A passive sensing node for monitoring the integrity of RF coaxial cables and connectors was described. The PSN is packaged on an MID and placed inside the RF connector body at the time of manufacture. A network of PSN devices was also described which provides a real-time, self-diagnosing, remotely addressable system for monitoring the integrity of RF coaxial transmission lines and associated RF connectors in the UHF range of frequencies. Each PSN includes three sensors which monitor connector tightness, moisture ingress, and temperature. The operation of these sensors was also described.

Finally, a compact model for the intra-coaxial RF Transmission line Coupler (TXC) was presented. The TXC is a crucial component of the PSN since it provides a means of harvesting power from the transmission line as well as a means of communicating the status of PSN sensors. This model expresses the spectral \( Z \) and \( S \) matrices in terms of fifteen model parameters. Five of these parameters can be computed from first principles and the remaining ten are obtained by fitting the model results to results obtained from ANSOFT’s numerical electromagnetic field simulator HFSS™. Agreement between model results and HFSS simulations was shown to be very good. Experimental measurements of the \( S \)-parameters of the RF coupler, power monitoring, power harvesting, and backscatter telemetry, along with a brief discussion on communication protocol for the PSN system are presented elsewhere [12].

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