Mathematical Modeling and Simulation of Aviation 1553B Bus Based on Parameter Identification

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Abstract. On the basis of the transmission line theory, combined with the characteristics of 1553B bus and its structures, distribution parameters of the equivalent differential circuit, the damage was deduced uniformly arbitrary input output response of the 1553B bus, propagation constant matrix, characteristic impedance matrix and transfer function matrix, established the mathematical model of cable frequency domain after cleverly simplified model, and with the test data in matlab by means of parameter identification, to estimate the parameters of cable, eventually have to close to the real value of nondestructive line mathematical models, and to the related parameters to build simulation model to do simulation experiment, the experimental results show the correctness of the mathematical model, It provides reference for the research of 1553B bus.

1. Introduction

The 1553B bus is a typical shielded twisted pair (STP). With its strong anti-interference, high reliability, flexibility and convenience, it is widely used in airborne / missile / shipborne integrated fire control systems, artificial satellites, etc. On military platforms, these different application carriers need to be flexible and convenient to accurately predict the transmission characteristics of the channel during the design process, thereby ensuring the quality of signal transmission. In addition, new applications have proposed high bandwidth resources and high communication rates for the 1553B bus system, which are suitable for long-distance and large-capacity fiber-optic communications. In addition, in order to solve the problems of multipath and crosstalk encountered during channel expansion Clearly understand channel characteristics. It can be seen that to realize the flexible design and application of the 1553B bus system and complete the improvement of the data transmission rate of the entire bus system, it is urgent to accurately describe the channel transmission characteristics of the 1553B bus system [1].

In this paper, Kirchhoff's law of transmission line theory- "path" method is used in combination with parameter identification to study the 1553B bus and its transmission characteristics [2]. With reference to the single-ended transmission line, a 1553B bus distributed parameter equivalent circuit model is established, and the output response, propagation constant matrix, characteristic impedance matrix, and transfer function matrix of any input that is lossy and uniform to the 1553B bus are derived based on this combination The sinusoidal observation data were used to identify the parameters of the transfer function matrix of its simplified lossless 1553B bus model, and the circuit model and mathematical
model modeling built on the matlab using the identified distribution parameters were simulated and verified respectively.

2. Transmission characteristics and output response of lossy uniformly shielded twisted pair wires

The 1553B bus belongs to the coupling multi-conductor transmission line system [3]. Different from the solitary conductor transmission line system, the conductors in the coupled multi-conductor system will interact with each other due to the action of electric field and magnetic field during signal transmission, resulting in crosstalk. According to the electromagnetic field theory, the distribution parameters of the uniform transmission line are evenly distributed along the cable, regardless of the position. The equivalent differential circuit model of the 1553B bus per unit length, as shown in figure 1, is built.

\[\begin{align*}
\{U_1(z) - U_1(z + dz) &= (R_0 + j\omega L_0)I_1(z)dz + j\omega M I_2(z)dz \\
\{U_2(z) - U_2(z + dz) &= (R_0 + j\omega L_0)I_2(z)dz + j\omega M I_1(z)dz \end{align*}\]  

(1)

Similarly, the basic equations of unit cable current

\[\begin{align*}
\{I_1(z) - I_1(z + dz) &= (G_0 + j\omega C_0)U_1(z + dz)dz + j\omega C M[U_1(z) - U_1(z + dz)]dz \\
\{I_2(z) - I_2(z + dz) &= (G_0 + j\omega C_0)U_2(z + dz)dz + j\omega C M[U_2(z) - U_2(z + dz)]dz \end{align*}\]  

(2)

Figure 1. Damage-shielded twisted pair equivalent differential circuit.

2.1. Transmission characteristics

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2.1.1. Transmission characteristics. As can be seen from the figure, because there are mutual inductance and mutual capacitance between two cables, the voltage on any cable depends not only on its own inductance and current, but also on the current of the other cable. According to the distribution parameter circuit model of cable and kirchhoff's law [4], the basic equations of unit cable voltage can be obtained:
Where $L_0$, $C_0$, $R_0$, $G_0$ represent the inductance, capacitance, resistance, and conductance distribution parameters of the twisted pair, and $L_M$ and $C_M$ respectively represent the mutual inductance and capacitance between the two wires, $U_1(z)$, $U_2(z)$, and $I_1(z)$, $I_2(z)$ represent the voltage and current of the two wires at $z$, the beginning of the twisted pair, and $U_1(z + dz)$, $U_2(z + dz)$, and $I_1(z + dz)$, $I_2(z + dz)$ represent the voltage and current after the passage of the $dz$ segment. The above equation takes the limit $dz \to 0$ and converts it into a matrix to obtain the first-order ordinary differential equation of shielded twisted pair, also known as the transmission line equation [5].

\[
\begin{align*}
\frac{dU}{dz} &= -ZI \\
\frac{dI}{dz} &= -YU
\end{align*}
\]  

(3)

By differentiating both sides of formula (3) and introducing a constant matrix of propagation $\gamma$, the wave equation of twisted pair can be shielded.

\[
\begin{align*}
\frac{d^2U}{dz^2} &= ZYU = \gamma^2U \\
\frac{d^2I}{dz^2} &= ZYI = \gamma^2I
\end{align*}
\]  

(4)

In the equation $U = [U_1 \ U_2]^T$, $I = [I_1 \ I_2]^T$. Where $Z = \begin{bmatrix} R_0 + j\omega L_0 & j\omega L_M \\ j\omega L_M & R_0 + j\omega L_0 \end{bmatrix}$ is the impedance matrix and $Y = \begin{bmatrix} G_0 + j\omega(C_0 + C_M) & j\omega C_M \\ j\omega C_M & G_0 + j\omega(C_0 + C_M) \end{bmatrix}$ is admittance matrix of the shielded twisted pair.

2.2. Output response

The degradation method is used to solve the voltage second-order homogeneous equations corresponding to the first term in formula (4)

\[
\begin{bmatrix} U_1'' \\ U_2'' \end{bmatrix} - ZY \begin{bmatrix} U_1 \\ U_2 \end{bmatrix} = \begin{bmatrix} 0 \\ 0 \end{bmatrix}
\]  

(5)

It can be found that $Z$, $Y$, $ZY$ are complex symmetric matrices, and it can be inferred that they are also complex symmetric matrices, and the main diagonal elements are the same.

Assume $\gamma = \begin{bmatrix} \gamma_1 \\ \gamma_2 \end{bmatrix}$. Equations are reduced to

\[
\begin{bmatrix} U_1'' \\ U_2'' \end{bmatrix} - \gamma^2 \begin{bmatrix} U_1 \\ U_2 \end{bmatrix} = \begin{bmatrix} 0 \\ 0 \end{bmatrix}
\]  

(6)

Reduce the order of the equation and do the transformation $U_1' = U_3$, $U_2' = U_4$. Substituted into the above formula

\[
\begin{bmatrix} U_1' \\ U_2' \\ U_3' \\ U_4' \end{bmatrix} - \begin{bmatrix} 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \\ \gamma_1 & \gamma_2 & 0 & 0 \\ \gamma_1 & 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} U_1 \\ U_2 \\ U_3 \\ U_4 \end{bmatrix} = \begin{bmatrix} 0 \\ 0 \\ 0 \\ 0 \end{bmatrix}
\]  

(7)

Using matrix analysis and calculus-related knowledge, the solution to the first-order homogeneous equation is
\[ U(z) = V(\exp(\Delta z) C_1 + \exp(-\Delta z)C_2) \]  

Where \( U = [U_1 \ U_2]^T \), \( \lambda = \text{diag}(\lambda_1, \lambda_2) = \begin{bmatrix} \lambda_1 + \lambda_2 & 0 \\ 0 & \lambda_1 + \lambda_2 \end{bmatrix} \), \( \lambda_1, \lambda_2 \) is the eigenvalue of matrix \( \gamma = \begin{bmatrix} y_1 \\ y_2 \\ y_1 \end{bmatrix} \), \( \mathbf{V} = \begin{bmatrix} 1 \\ 1 \end{bmatrix} \) is the column eigenvector of matrix \( \gamma = \begin{bmatrix} y_1 \\ y_2 \\ y_1 \end{bmatrix} \). \( C_1, C_2 \) is a \( 2 \times 1 \) Constant matrix. Similarly, \( I(z) = (V(\exp(\Delta z) C_1 + \exp(-\Delta z)C_2)/Z_0) \).

Where \( Z_0 \) is the stable value of characteristic impedance, \( U \) and \( I \) are \( 2 \times 1 \) order matrix functions about distance \( z \), \( Z \) is the unit of length. In order to express the output response of the shielded twisted pair more directly, \( l \) use instead of \( z \), so the voltage and current output response of the shielded twisted pair of length \( l \) is

\[
\begin{align*}
&\begin{align*}
&\{U(l) = V(\exp(\Delta l) C_1 + \exp(-\Delta l)C_2) \\
&I(l) = (V(\exp(\Delta l) C_1 + \exp(-\Delta l)C_2)/Z_0)
\end{align*}
\end{align*}
\]

making \( V = P, C_1 = P^{-1}a, C_2 = P^{-1}b \), Simplified

\[
\begin{align*}
&\begin{align*}
&\{U(l) = Pe^{\Delta l}P^{-1}a + Pe^{-\Delta l}P^{-1}b \\
&I(l) = Pe^{\Delta l}P^{-1}a/Z_0 + Pe^{-\Delta l}P^{-1}b/Z_0
\end{align*}
\end{align*}
\]

3. Shielded twisted pair characteristic impedance matrix and transfer function matrix

3.1. Characteristic impedance matrix

According to the idea of finite element analysis, a continuous transmission line can be approximated as \( n \) unit length transmission line. If the input impedance of an infinite number of small cascade networks is \( Z_{C} \), cascade a small section before this network to obtain the total network and its input. Impedance remains the same [7]. The unit equivalent length shielded twisted pair distributed parameter equivalent circuit in Figure 1 is generalized, and a lossy shielded twisted pair discrete equivalent circuit model shown in Figure 2 can be established.

![Figure 2. Discrete equivalent circuit model of lossy shielded twisted pair](image)

\( Z_{C} \) represents the input impedance of the infinitely many discrete small-stage front cascade networks that are calculated later, and it is slightly different from the characteristic impedance of continuous transmission lines. It is mathematically described as cascading a small segment before the infinitely
small cascade network, paralleling the input impedance of the network \( \overline{Z}_C \) with the parallel admittance \( Y \), and then adding it to the series impedance matrix \( Z \), the total input impedance is still

\[
\overline{Z}_C = Z + \frac{l}{i/\overline{Z}_C + Y}
\]

(11)

Where \( \overline{Z}_C, Z, Y \) are 2x2 complex symmetric square matrices and \( I \) are second-order identity matrices. Simplified

\[
\overline{Z}_C = \frac{Z}{\sqrt{Y} + Z\overline{Z}_C}
\]

(12)

The above formula is the input impedance of an infinite number of discrete small lumped parameter cascade networks. Of course, this discrete small segment lumped parameter cascade network is only an approximation of the continuous transmission line structure, but the shorter the size (length) of each small segment, the more the number of cascade segments, the better the approximation. This kind of discrete network model can accurately describe the continuous transmission line structure.

The distribution parameters of each segment in Figure 2 are \( R_0, L_0, C_0, G_0, C_M, L_M \). If each segment is further subdivided into a cascade of segments[6], after segmentation, the distribution parameters of the smallest segment become \( R_0/n, L_0/n, C_0/n, G_0/n, C_M/n, L_M/n \). The series impedance matrix \( Z \) and the parallel admittance matrix \( Y \) become \( Z/n \) and \( Y/n \). Substituting these new parameters into Eq. (12) and taking the limit towards infinity, the input impedance of the continuous transmission line can be obtained. When \( n \) takes the limit towards infinity, it tends to 0, and the determined expression \( Z_C \) obtained at this time

\[
Z_C = \sqrt{Z/Y} = \sqrt{Z_Y^{-1}}
\]

(13)

Where \( Z = \begin{bmatrix} R_0 + j\omega L_0 & j\omega L_M \\ j\omega L_M & R_0 + j\omega L_0 \end{bmatrix} \), \( Y = \begin{bmatrix} G_0 + j\omega(C_0 + C_M) & j\omega C_M \\ j\omega C_M & G_0 + j\omega(C_0 + C_M) \end{bmatrix} \) is the impedance matrix and the admittance matrix.

In general, it changes significantly with frequency. In order to accurately calculate the characteristic impedance matrix \( Z_C \), Decomposing \( Y \) into

\[
Y = A + jB
\]

(14)

Where \( A = \begin{bmatrix} G_0 & 0 \\ 0 & G_0 \end{bmatrix}, B = \begin{bmatrix} \omega(C_0 + C_M) & \omega C_M \\ \omega C_M & \omega(C_0 + C_M) \end{bmatrix} \), Where the real part matrix \( A \) and the imaginary part matrix \( B \) are Second-order invertible real matrix. Due to \( (AB^{-1} + BA^{-1})(AB^{-1} + BA^{-1})^{-1} = (A + jB)(B^{-1} - jA^{-1})(AB^{-1} + BA^{-1})^{-1} \). From the matrix inversion formula \( (B^{-1} - jA^{-1})(AB^{-1} + BA^{-1})^{-1} = (A + BA^{-1}B)^{-1} - jA^{-1}B(A + BA^{-1}B)^{-1} \). The inverse matrix of the parallel admittance matrix is

\[
Y^{-1} = (A + BA^{-1}B)^{-1} - jA^{-1}B(A + BA^{-1}B)^{-1}
\]

(15)

Where \( A^{-1} = \begin{bmatrix} 1/G_0 & 0 \\ 0 & 1/G_0 \end{bmatrix}, B^{-1} = \begin{bmatrix} \omega C_M & \omega C_M \\ \omega C_M & \omega C_M \end{bmatrix} \), However \( Z_C = \sqrt{Z_Y^{-1}} \), calculating and simplifying...
The length of each twisted pairs. Each discrete segment of unit length can still be further divided into small segments. In summary, the independent cascade of discrete segments is only an approximation of continuous twisted pairs. Each discrete segment of unit length can still be further divided into n small segments. The length of each segment is 1/n, the series impedance matrix is Z/n, and the parallel admittance matrix is Y/n. The n-stage cascaded response is equal to the nth power of the sub-segment response of

\[
Z_C = \begin{bmatrix}
\frac{\sqrt{(Z_X + \sqrt{Z_Y})/2}}{Z} \\
\frac{Z - \sqrt{(Z_X - \sqrt{Z_Y})/2}}{Z} \\
\end{bmatrix}
\]

(16)

Where \( Z_X = ((\omega^2(C_0 + C_M)^2 + \omega^2 C_M^2 + G_0^2)(R_0 G_0 + \omega^2 L_0 (C_0 + C_M) + j \omega (L_0 C_M - R_0 (C_0 + C_M))) + 2 \omega^2 C_M L_M (C_0 + C_M) (-\omega^2 C_M + j \omega G_0)) \), \( Z_Y = [((R_0 G_0 + \omega^2 (L_0 + L_M) (C_0 + C_M)) + j \omega (L_0 C_M + L_M G_0 - R_0 (C_0 + C_M))) + 2 \omega^2 C_M (C_0 + C_M) (R_0 G_0 - 2 \omega^2 L_M C_M + j \omega (R_0 C_M + 2 L_M G_0))][(\omega^2 (C_0 + C_M) + \omega^2 C_M^2 + G_0^2) (R_0 G_0 + \omega^2 (L_0 - L_M) (C_0 + C_M) + j \omega (L_0 C_M - L_M G_0 - R_0 (C_0 + C_M))) - 2 \omega^2 L_M R_0 (C_0 + C_M) (G_0 + j \omega C_M)] \), \( \hat{Z} = (\omega^2 (C_0 + C_M) + \omega^2 C_M^2 + G_0^2) + 2 j \omega^2 C_M (C_0 + C_M) \).

3.2. Transfer function matrix and transfer constant matrix

When deriving the transfer function matrix model, a strategy of adding a discrete segment of unit length before a continuous shielded twisted pair transmission line with an input impedance matrix of is adopted. As shown in Figure

Figure 3. impedance distribution of continuous shielded twisted pair.

If \( Z' \) is used to represent the input impedance matrix at the right end of line A, according to the voltage distribution principle, the transmission constant \( \bar{H} \) is

\[
\bar{H} = \frac{Z'}{Z + Z'}
\]

(17)

In the above formula, it is equal to the parallel connection of the admittance matrix \( Y \) and the input impedance \( Z_C \), where \( Z' = \frac{1}{Y + 1/Z_C} \), \( Z_C = \sqrt{YZ^{-1}} \), substituting it into formula (17) Simplified

\[
\bar{H} = \frac{1}{2Y + \sqrt{2Y + 1}}
\]

(18)

Equation (18) is the transfer function of the shielded twisted-pair cable with discrete length per unit length.
length $1/n$. Sum the transfer functions of the shielded twisted-pair wires of discrete segments to obtain the transfer function of the continuous shielded twisted-pair wires.

$$ H = \lim_{n \to \infty} \left[ \frac{1}{\left( \frac{1}{n} \right)^2 + \sqrt{\frac{1}{n} + \frac{1}{n}} + 1} \right]^n \quad (19) $$

According to the mathematical formula below $\lim_{n \to \infty} \left( 1 + \frac{a}{n} \right)^n = e^{-a}$, the above formula can be simplified

$$ H = \lim_{n \to \infty} \left( 1 + \frac{ZY}{n} + \frac{\sqrt{ZY}}{n} \right)^{-\frac{n}{\sqrt{ZY} + \sqrt{ZY}}} = e^{-\lim_{n \to \infty} \left( \sqrt{ZY} + \sqrt{ZY} / \pi \right)} \quad (20) $$

Considering that $n$ in Eq. (20) takes the limit towards infinity, $ZY/n$ tends to zero, and we get

$$ H = e^{-\sqrt{ZY}} \quad (21) $$

Where $Z = \left[ \begin{array}{cc} R_0 + j\omega L_0 & j\omega M \\ j\omega L_0 & R_0 + j\omega L_0 \end{array} \right], Y = \left[ \begin{array}{cc} G_0 + j\omega (C_0 + C_M) & j\omega C_M \\ j\omega C_M & G_0 + j\omega (C_0 + C_M) \end{array} \right]$

From Equation (21), the propagation constant matrix can be obtained $\gamma(\omega) = \sqrt{ZY}$. The transmission function matrix of the lossy shielded twisted pair per unit length can be expressed as $H(\omega) = e^{-\gamma(\omega)}$, Abbreviated as $H = e^{-\gamma}$, so there is $\gamma(\omega) \equiv -\ln(H(\omega))$.

In the above definition of the formula $\gamma$, a negative sign is added in front. This is because the real part of $\gamma$ is a positive value, which is consistent with the concept that the transmission line is always always attenuating. The natural logarithm (plus minus sign) of the shielded twisted pair transmission function matrix $H$ of unit length is called the propagation constant matrix, and the unit is complex $Np / m$.

From $Z$, $Y$, $ZY$ is a $2 \times 2$ complex symmetric matrix, which can be deduced to be a complex symmetric matrix with the same diagonal elements. Assume $\gamma = \left[ \begin{array}{cc} \gamma_1 & \gamma_2 \\ \gamma_2 & \gamma_1 \end{array} \right]$, Where $\gamma_1$, $\gamma_2$ both are variables about frequency. The propagation constant of a single wire of a twisted wire is caused by the electromagnetic field of another wire to this wire. Because $\gamma = \sqrt{ZY}$ and $\gamma_1 \geq \gamma_2$. Simplified by the undetermined coefficient method

$$ \gamma_1 = \sqrt{\frac{\gamma_1 + \sqrt{\gamma_2}}{2}}, \gamma_2 = \sqrt{\frac{\gamma_1 - \sqrt{\gamma_2}}{2}} \quad (22) $$

Where $\gamma_x = \omega^2 L_M C_M + R_0 G_0 - \omega^2 L_0 (C_0 + C_M) + j\omega (L_0 G_0 + R_0 (C_0 + C_M))$ and $\gamma_y = \left[ (R_0 G_0 + \omega^2 L_0 (C_0 + C_M) + j\omega (L_0 G_0 + R_0 (C_0 + C_M)) \right] \left[ (R_0 G_0 + \omega^2 (2C_M - C_0) (L_0 - L_M)) + j\omega (G_0 (L_M - L_0) + R_0 (C_0 + 2C_M)) \right]$

From the above, the transfer function matrix $H$ is a state transition matrix of the function $f(t) = e^{-t}$. By analyzing the relevant knowledge of the linear system, the transfer function matrix can be obtained

$$ H = f(\gamma) = Pf(\gamma)P^{-1} \quad (23) $$

Where $J = \left[ \begin{array}{cc} \gamma_1 + \gamma_2 & 0 \\ 0 & \gamma_1 + \gamma_2 \end{array} \right]$ is a Jordanian type of matrix $\gamma(\omega)$, $P = \left[ \begin{array}{cc} 1 & -1 \\ 1 & 1 \end{array} \right]$, $P^{-1} = \left[ \begin{array}{cc} 1/2 & 1/2 \\ 1/2 & 1/2 \end{array} \right]$ is the transformation matrix and its inverse matrix obtained from the feature vector
corresponding to the feature vector equation $AP_i = \lambda P_i (i = 1, 2)$, and $f(f) = \begin{bmatrix} f(y_1 + y_2) \\ 0 \end{bmatrix}$ is $J$ state transition matrix for $f(t) = e^{-t}$. The transfer function matrix can be obtained as

$$H = \begin{bmatrix} e^{-\gamma_1} \left( e^{\gamma_2} + e^{-\gamma_2} \right) \frac{z}{2} & -e^{-\gamma_1} \left( e^{\gamma_2} - e^{-\gamma_2} \right) \frac{z}{2} \\ -e^{-\gamma_1} \left( e^{\gamma_2} - e^{-\gamma_2} \right) \frac{z}{2} & e^{-\gamma_1} \left( e^{\gamma_2} + e^{-\gamma_2} \right) \frac{z}{2} \end{bmatrix} \quad (24)$$

Definition of hyperbolic sine function and hyperbolic cosine function $\sinh(x) = \frac{1}{2} (e^x - e^{-x})$, $\cosh(x) = \frac{1}{2} (e^x + e^{-x})$. The unit length shielded twisted pair transmission function is arrayed as

$$H = \begin{bmatrix} e^{-\gamma_1 \cosh (\gamma_2)} & -e^{-\gamma_1 \sinh (\gamma_2)} \\ -e^{-\gamma_1 \sinh (\gamma_2)} & e^{-\gamma_1 \cosh (\gamma_2)} \end{bmatrix} \quad (25)$$

According to the transmission line theory, when a signal is transmitted in a twisted pair, the signal amplitude decreases exponentially with the transmission distance by a certain factor $H$. However, this attenuation factor $H$ is the same for each segment and varies with frequency. Therefore, $H(\omega)$ let the change in frequency of the twisted-pair wire per unit length be called the transfer function matrix of the twisted pair wire. $H(\omega, l)$ indicates the change of the twisted pair with a length of $l$ relative to the frequency, then $H(\omega, l)$ and $H(\omega)$ have the following exponential relationship: $H(\omega, l) = [H(\omega)]^l$, that is, a shielded twisted pair with a certain length of $l$ can be equivalent to an infinite number of consecutive small unit twisted pairs. Cascaded network [8]. so

$$H(\omega, l) = \left[ e^{-\gamma(\omega)} \right]^l = e^{-\gamma(\omega) l}$$

(26)

Where $l$ is the length of the shielded twisted pair, the unit is m; $\gamma(\omega)$ is the propagation constant matrix of the shielded twisted pair at the frequency (unit is rad / s) (the unit is complex Np / m). So the transfer function matrix of a certain length of lossy shielded twisted pair is

$$H(\omega, l) = \begin{bmatrix} e^{-\gamma_1 l \cosh (\gamma_2 l)} & -e^{-\gamma_1 l \sinh (\gamma_2 l)} \\ -e^{-\gamma_1 l \sinh (\gamma_2 l)} & e^{-\gamma_1 l \cosh (\gamma_2 l)} \end{bmatrix} \quad (27)$$

3.3. Lossless shielded twisted pair transfer function array
The lossless twisted pair is a simplification of the lossy twisted pair. At high frequencies, the resistance and conductance are omitted. The lossy wire model is simplified as the lossless shielded twisted pair model. As shown Figure 4.
In order to identify the need, when calculating the mathematical model of the lossless line, at high frequencies, the resistance $R_0$ and conductance $G_0$ are ignored, and the coefficient of the propagation constant matrix $\gamma_1$, $\gamma_2$ of the lossy transmission line is converted into an $s$-domain expression using $s = j\omega$, $\gamma_1$, $\gamma_2$ can be simplified as

$$
\gamma_1' = s\sqrt{\frac{\gamma_x'}{2}}, \quad \gamma_2' = s\sqrt{\frac{\gamma_y'}{2}}.
$$

(28)

Where $\gamma_x' = L_0 C_0 + (L_0 - L_M) C_M$, $\gamma_y' = (L_0^2 - L_M^2)(C_0 + 2C_M) C_0$.

Therefore, the transmission function matrix of the lossless shielded twisted pair per unit length is

$$
H = \begin{bmatrix}
e^{-\gamma_1' \frac{1}{2} (e^{\gamma_x'} + e^{-\gamma_y'})} & -e^{-\gamma_1' \frac{1}{2} (e^{\gamma_x'} - e^{-\gamma_y'})} \\
e^{-\gamma_1' \frac{1}{2} (e^{\gamma_y'} - e^{-\gamma_x'})} & e^{-\gamma_1' \frac{1}{2} (e^{\gamma_y'} + e^{-\gamma_x'})}
\end{bmatrix}
$$

(29)

The transfer function matrix of a lossless shielded twisted pair with a certain length $l$ is

$$
H(\omega, l)' = \begin{bmatrix}
e^{-\gamma_1' \frac{l}{2} (e^{\gamma_x'} + e^{-\gamma_y'})} & -e^{-\gamma_1' \frac{l}{2} (e^{\gamma_x'} - e^{-\gamma_y'})} \\
e^{-\gamma_1' \frac{l}{2} (e^{\gamma_y'} - e^{-\gamma_x'})} & e^{-\gamma_1' \frac{l}{2} (e^{\gamma_y'} + e^{-\gamma_x'})}
\end{bmatrix}
$$

(30)

4. Experiments and results

4.1. Lossless Shielded Twisted Pair Transfer Function Matrix Model and Parameter Identification

For the convenience of research and identification, the distribution parameters $L_0$, $C_0$, $C_M$, $L_M$ of non-destructive shielded twisted-pair cables are calculated, and are processed as follows. In the entire frequency range, no matter which mode is used, the transmission line can always be shortened to a certain length, so that it is not It works in distributed parameter mode, but works in lumped parameter circuit state.

Referring to a single-ended transmission line, since the amplitude of the propagation constant $l\gamma$ is limited in the lumped parameter characteristic region, in this region, Using a finger function of $e$ to
shorten \( H_{Y_1}'(\omega) = e^{-\gamma_1'(\omega)} \), \( H_{Y_2}'(\omega) = e^{-\gamma_2'(\omega)} \), \( H_{Y_1}'^{-1} = e^{\gamma_1'(\omega)} \) and \( H_{Y_2}'^{-1} = e^{\gamma_2'(\omega)} \) using Taylor series expansion can greatly simplify the analysis process: \( H_{Y_1}' = 1 - \gamma_1' + \frac{\gamma_1'^2}{2!} \ldots \), \( H_{Y_1}'^{-1} = 1 + \gamma_1' + \frac{\gamma_1'^2}{2!} + \frac{\gamma_1'^3}{3!} \ldots \), \( H_{Y_2}' = 1 - \gamma_2' + \frac{\gamma_2'^2}{2!} - \frac{\gamma_2'^3}{3!} \ldots \), \( H_{Y_2}'^{-1} = 1 + \gamma_2' + \frac{\gamma_2'^2}{2!} + \frac{\gamma_2'^3}{3!} \ldots \).

Take the first four terms of the above expressions and add and subtract them, respectively, to get the approximate formula used to calculate the transfer function matrix of the transmission line:

\[
\frac{H_{Y_1} + H_{Y_1}^{-1}}{2} \approx 1 + \frac{\gamma_1'^2}{2!} + \frac{\gamma_1'^3}{3!} \approx \gamma_1' + \gamma_1'^3 \quad (31)
\]

Due to \( H_{Y_1}' = \frac{1}{H_{Y_1}'^{-1}} \approx \frac{1}{1 + \gamma_1' + \gamma_1'^2 + \gamma_1'^3} \quad (32) \)

So the approximate expression of the transfer function matrix for unit length twisted pairs is

\[
H = \left[ \begin{array}{c}
\left( \frac{1}{H_{Y_1}'^{-1}} \right) \left( \frac{H_{Y_1} + H_{Y_1}^{-1}}{2} \right) - \left( \frac{1}{H_{Y_1}'^{-1}} \right) \left( \frac{H_{Y_2} + H_{Y_2}^{-1}}{2} \right) \\
\left( \frac{1}{H_{Y_1}'^{-1}} \right) \left( \frac{H_{Y_2} + H_{Y_2}^{-1}}{2} \right) - \left( \frac{1}{H_{Y_1}'^{-1}} \right) \left( \frac{H_{Y_1} + H_{Y_1}^{-1}}{2} \right)
\end{array} \right]
\]

(33)

Let \( \gamma_1' = \mu \), \( \gamma_2' = \lambda s \) the lossy line transfer function unitization of unit length is

\[
H = \left[ \begin{array}{c}
\frac{1 + \mu^2 s^2}{2} \frac{\lambda^2 s^3}{6} \\
\frac{1 + \mu^2 s^2}{2} \frac{\lambda^2 s^3}{6}
\end{array} \right]
\]

(34)

Where \( \mu = \sqrt{\frac{L_0 C_0 + (L_0 - L_M) C_M + (L_0 - L_M)^2 (C_0 + C_M) C_0}{2}} \), \( \lambda = \sqrt{\frac{L_0 C_0 + (L_0 - L_M) C_M + (L_0 - L_M)^2 (C_0 + C_M) C_0}{2}} \).

It can be found that the elements in the transfer function matrix correspond to the ratio of the output and input of the corresponding branch. The ratio of the cross output to the input is the transfer function of the coupling term between the branches. It also fits the two wires of the twisted pair to each other. Characteristics. The system described by the transfer function matrix is a two-input two-output linear invariant system, and the distribution parameters of the two wires of the twisted pair are the same. Therefore, it is possible to carry out parameter identification research on the transfer function under the condition of measuring input and output. Set the input and output as \( u = [u_1 \ u_2]^T \), \( y = [y_1 \ y_2]^T \) the system can be decoupled under common mode input (that is, when \( u_2 = 0 \)), and the identification system can be split into

\[
\begin{align*}
A(s) y_1(s) &= B(s) u_1(s) \\
C(s) y_2(s) &= D(s) u_1(s)
\end{align*}
\]

(35)

Where \( A(s) = 1 + \mu s + \frac{\mu^2 s^2}{2} + \frac{\mu^3 s^3}{6}, B(s) = 1 + \frac{\lambda^2 s^2}{2}, C(s) = -A(s), D(s) = \lambda s + \frac{\lambda^3 s^3}{6} \).

Transformed into differential equations and simplified by discrete difference
\[
\begin{align*}
y_1(k) + a_1 y_1(k-1) + a_2 y_1(k-2) + a_3 y_1(k-3) = b_0 u_1(k) + b_1 u_1(k-1) + b_2 u_1(k-2) \\
y_2(k) + a'_1 y_2(k-1) + a'_2 y_2(k-2) + a'_3 y_2(k-3) = b'_0 u_1(k) + b'_1 u_1(k-1) + b'_2 u_1(k-2)
\end{align*}
\] (36)

Where \( a_1 = a'_1 = -\frac{\mu^2 + \mu^3}{\mu + \mu^2 + \mu^3}, \quad a_2 = a'_2 = \frac{\mu^2 + \mu^3}{\mu + \mu^2 + \mu^3}, \quad a_3 = a'_3 = \frac{\mu^3}{\mu + \mu^2 + \mu^3}, \quad b_0 = \frac{1 + \mu^2}{\mu + \mu^2 + \mu^3}, \quad b_1 = \frac{-\mu^3}{\mu + \mu^2 + \mu^3}, \quad b_2 = \frac{-\mu^3}{\mu + \mu^2 + \mu^3}, \quad b'_0 = \frac{1 + \mu^2}{\mu + \mu^2 + \mu^3}, \quad b'_1 = \frac{-\mu^3}{\mu + \mu^2 + \mu^3}, \quad b'_2 = \frac{-\mu^3}{\mu + \mu^2 + \mu^3}.

The difference equation model is an ARMAX model, so the identification model is an ARMAX model. The formula (14) is transformed into the least squares format.

\[
y_i(k) = h_i^T(k) \theta_i + e_i(k), (i = 1,2)
\] (37)

Where \( h_1^T(k) = [y_1(k-1) y_1(k-2) y_1(k-3) u_1(k) u_1(k-1) u_1(k-2)] \), \( h_2^T(k) = [y_2(k-1) y_2(k-2) y_2(k-3) u_1(k) u_1(k-1) u_1(k-2) u_1(k-3)] \). \( \theta_1 = [a_1 a_2 a_3 b_0 b_1 b_2]^T \), \( \theta_2 = [a'_1 a'_2 a'_3 b'_0 b'_1 b'_2]^T \). \( y_i(k) \) and \( h_i^T(k) \) is Observable data. \( \theta_i \) is the parameter to be identified, \( e_i(k) \) is the model and real output error. The selected objective function is

\[
J_i(\theta) = \sum_{k=1}^{L} [e_i(k)]^2 = \sum_{k=1}^{L} [y_i(k) - h_i^T(k) \theta_i]^2 (i = 1,2)
\] (38)

Where \( L \) is the length of the observable data.

4.2. Experiment

The experimental setup is shown in Figure 5.

![Experimental system diagram](image)

**Figure 5.** Experimental system diagram

The signal frequency of the 1553B bus is 1MHz, and the sampling period of the signal acquisition card is 5 ~ 10us. In the laboratory, the signal generator injects a common-mode excitation signal into the shielded twisted pair (that is, one end injects the excitation signal and one end does not). After the signal passes through the 1553B bus, each of the positive and negative electrodes at the output end outputs a signal after passing through the signal conditioning board and the input excitation. The signals are sent to the signal acquisition card together, and then the data is collected by the signal acquisition card and uploaded to the computer. Finally, the parameter identification and calculation are completed on the matlab software [9].
### Table 1. 1553B bus model and parameters.

| Component       | Model              | Subhead                                                                 |
|-----------------|--------------------|-------------------------------------------------------------------------|
| Twisted pair    | M17/176-00002      | Characteristic impedance 77 ± 7Ω, standard capacitance 24pF / ft, attenuation 1.4dB / 100ft at 1MHz, insulation material Teflon, center conductor diameter 0.0235mm, insulator diameter 0.042mm, shield diameter 0.1mm, total wire diameter 0.129mm |

4.3. *Input excitation signal*

In order to identify the distribution parameters of the 1553B bus, it is necessary to inject a "full excitation signal" into the 1553B bus. The so-called "excitation is sufficient" means that the excitation signal contains sufficient frequency components and has a certain excitation time. Since the single-frequency sinusoidal signal is a second-order continuous excitation signal, it can only identify two systems with unknown parameters. If the number of identification parameters is \( n \), for this purpose, at least \( 2n \) sinusoidal signals combined with sinusoidal signals of different frequencies are used as excitation signals to generate observation data for parameter identification. If the number of different frequencies of the multi-frequency sinusoidal combined signal is \( m \), then \( m \geq n \) is required. Sine combined signals of different frequencies can be expressed as

\[
 u(t) = \sum_{i=1}^{m} a_i \sin (\omega_i t) \tag{39}
\]

Where \( a_i \) is the amplitude of each sine component and \( \omega_i \) is the angular frequency of each sine component.

In the identification experiment, the amplitude and phase are artificially set and are known. To meet the conditions of sufficient excitation, the injection input excitation signal is selected as

\[
 u(t) = \sin(2\pi \times 10^6 t) + \sin(4\pi \times 10^6 t) + \sin(6\pi \times 10^6 t) + \sin(8\pi \times 10^6 t) \tag{40}
\]

![Identifying experimental stimulus signals](image)

**Figure 6.** Identifying experimental stimulus signals

In this way, the identification conditions of the lossy twisted pair transmission function are satisfied, and the parameter identification of the distribution parameters of the lossy twisted pair can be performed.
4.4. Experimental results

The parameters were identified by least squares and genetic algorithms. The genetic algorithm population size is 300, the number of evolutionary generations is 600, the crossover frequency is 0.90, and the mutation frequency is 0.01. The distribution parameters of the identification results after solving are shown in Table 2. Although the least square method is also used for identification, the accuracy is too low, and the result of the genetic algorithm is relatively close to the estimated value.

| Distribution parameter | Results         |
|------------------------|-----------------|
| $C_0$                  | 62.80(pF)       |
| $L_0$                  | 304.65(nH)      |
| $C_M$                  | 10.28(pF)       |
| $L_M$                  | 53.64(nH)       |

Table 2: Calculation results.

The transfer function matrix of the lossy shielded twisted pair is

$$
H(s) = \begin{bmatrix}
    e^{-1.4773 \times 10^{-8} s} \cosh(1.2049 \times 10^{-9} s) & -e^{-1.4773 \times 10^{-8} s} \sinh(1.2049 \times 10^{-9} s) \\
    -e^{-1.4773 \times 10^{-8} s} \sinh(1.2049 \times 10^{-9} s) & e^{-1.4773 \times 10^{-8} s} \cosh(1.2049 \times 10^{-9} s)
\end{bmatrix}
$$

(41)

Since the true values of the system distribution parameters are unknown, the accuracy of the identification parameters is tested. By giving the same stimulus signals to the real system and the circuit model and mathematical model established in matlab, observe their output results. To illustrate the general applicability of the model. Differential sine excitations at frequencies of 1MHz, 2 MHz, and 4MHz are given in the experiment. See the sine response of the model built with simulink and the real system.

Response 1 Differential sine excitations at frequencies of 1MHz

![Figure 7. Response curve of 1553B bus under 1MHz excitation signal](image-url)
Response 2 Differential sine excitations at frequencies of 2MHz

Figure 8. electronic circuit model diagram

Figure 9. mathematical model simulation diagram

Figure 10. Response curve of 1553B bus under 2MHz excitation signal
Figure 11. electronic circuit model diagram

Figure 12. mathematical model simulation diagram

Response 3  Differential sine excitations at frequencies of 4MHz

Figure 13. Response curve of 1553B bus under 4MHz excitation signal
From the experimental results, the output of the mathematical model is almost the same as the real output of the system, and it also shows the better delay characteristics of the 1553B bus with an accuracy of <3%. To some extent, the mathematical model can be considered to be equivalent to the real system. It also proves the correctness of the model.

5. Conclusion
This paper studies the equivalent differential circuit of the physical layer cable of the 1553B bus, and establishes numerical models such as its arbitrary input and output response, characteristic impedance matrix, propagation constant matrix, and transfer function matrix, and mathematically models the 1553B bus system through parameter identification. And mathematical methods are used to simplify the parameter identification process. Finally, the model is verified by simulation. The experimental results show the correctness of the model, verify the feasibility of the modeling method proposed in this paper, and provide relevant references for the study of modeling and simulation of 1553B.

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