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Errata to IEEE Recommended Practice for the Usage of Terms Commonly Employed in the Field of Large-Signal Vector Network Analysis

Developed by the

Standards Coordinating Committee

of the

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Correction Sheet
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STANDARDS

In 3.2.1 replace "y" in NOTE 3, NOTE 4, and Equation (5) as follows:

NOTE 3—Again for the linear sinusoidal steady state case, to be consistent with *non-TEM* transmission lines (e.g., rectangular waveguide), the transmission-line equivalent-circuit parameters should be defined from both the propagation constant γ that derives from the Maxwell electromagnetic field equations, and the characteristic impedance Z_0 of that transmission line determined by the chosen circuit theory with Equation(2) and Equation (3):

NOTE 4—When the per-unit-length equivalent-circuit parameters are already known, Z_0 and γ can be computed as shown in the following Equation (4) and Equation (5):

$$\gamma = \sqrt{ZY} = \sqrt{\left(R + j\omega L\right)\left(G + j\omega C\right)} \tag{5}$$

In 3.2.2 italicize "Z_0" in NOTE 3 as follows:

NOTE 3—Traveling waves in real transmission lines have some loss associated with them. This loss causes the traveling wave to slowly decay in the direction of propagation, and will change the phase relationships between the modal electric and magnetic fields carried by the traveling wave. This usually results in a complex modal characteristic impedance Z_0 [B3]. The imaginary part of Z_0 can usually be ignored in low-loss coaxial and rectangular-waveguide transmission lines. The imaginary part of Z_0 can be large in printed transmission lines, particularly at low frequencies.

In 3.2.4 revise Equation (6), Equation (7), and Equation (8) as follows:

$$a_p\left(Z_{ref}\right) := \frac{V_p + Z_{ref} I_p}{2\sqrt{Z_{ref}}},\ b_p\left(Z_{ref}\right) := \frac{V_p - Z_{ref} I_p}{2\sqrt{Z_{ref}}} \tag{6}$$

$$a_{RMS}\left(Z_{ref}\right) := \frac{V_{RMS} + Z_{ref} I_{RMS}}{2\sqrt{Z_{ref}}} = \frac{a_p}{\sqrt{2}},\ b_{RMS}\left(Z_{ref}\right) := \frac{V_{RMS} - Z_{ref} I_{RMS}}{2\sqrt{Z_{ref}}} = \frac{b_p}{\sqrt{2}} \tag{7}$$

$$a_{vw} := \sqrt{Z_{ref}}\, a_p = \frac{1}{2}\left(V_p + Z_{ref} I_p\right),\ b_{vw} := \sqrt{Z_{ref}}\, b_p = \frac{1}{2}\left(V_p - Z_{ref} I_p\right) \tag{8}$$

In 3.2.4 revise NOTE 4 to include plus/minus notation as follows:

NOTE 4—Defining $V_{\pm}$ (see traveling waves) as the peak voltages carried by the forward and backward traveling wave, the following equivalent normalization relations hold:

In 3.2.5 remove conjugation symbol in Equation (12) to appear as follows:

$$\hat{a}\left(\hat{Z}\right) = \frac{v + i\hat{Z}}{2\sqrt{\mathrm{Re}\left(\hat{Z}\right)}} \tag{12}$$

In 3.2.7 revise the variable in the "where" list after Equation (20) as follows:

where

$a_{vw} := \sqrt{Z_{ref}}\,a$ for the forward wave

In 3.2.8 revise the "where" list under Equation (23) as follows:

where

$a_{vw} := \sqrt{Z_{ref}}\,a_p$ for the forward wave

$b_{vw} := \sqrt{Z_{ref}}\,b_p$ for the backward wave

In 3.4.1 revise the text, Equation (24), and Figure 2 caption as follows:

3.4.1 Admissible pseudo-wave vector pair

An admissible pseudo-wave vector pair is any vector pair $(\mathbf{a}(t), \mathbf{b}(t))$, $t \geq t_0$, where t represents time and t_0 is a fixed demarcation time, of pseudo-waves associated to an N-port $\mathcal{N}$ (see Figure 2)—with the following Equation (24):

$$\mathbf{c}(t) := \begin{bmatrix} c_1(t) \\ \vdots \\ c_N(t) \end{bmatrix}, c = a, b \tag{24}$$

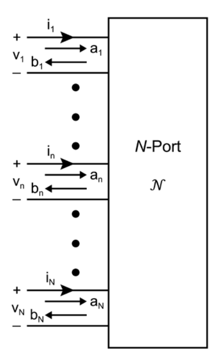

Figure 2—*N*-port $\mathcal{N}$

NOTE 1—The abstract concept of admissible pairs was originally couched in the lumped port voltage and current domain (see, for example, [B10]), and is the basis for the definition of a *constitutive relation* wherein the infinite set of admissible pairs can be reduced to some compact analytical/numerical/computational formula that relates the members of the pair (for example, Ohm's Law for a linear 2-terminal resistor that can be considered a 1-port). In addition, the port voltage and current pairing can be replaced with a more general circuit quantity pairing $\left(v^{(\alpha)}, i^{(\beta)}\right)$, where α and β are integers that represent the order of differentiation when positive, integration when negative, and neither when zero (that is, $x^{(0)} = x$, $x = v$, i). Thus, for example, a 2-terminal capacitor would constitute the pairing $\left(v^{(1)}, i^{(0)} = i\right)$ with the algebraic constitutive relationship $i = Cv^{(1)}$ if it were linear. In this case, the *N*-port is said to be *algebraic*, and is further called *mixed algebraic* if the exponents α_n, β_n are not identical across the *N*-ports. A yet further generalization of an algebraic *N*-port, termed a *dynamic N*-port, can be made if one or more of the ports requires more than two independent circuit variables $v^{(\alpha)}$ and $i^{(\beta)}$ to describe its constitutive behavior. An example of this would be a hysteretic 2-terminal element with the constitutive relation $v = g\left(i - f\left(v^{(-1)}\right)\right)$, thus involving $v^{(0)}, i^{(0)}$ and $v^{(-1)}$. See [B10] for a general and exhaustive set of details concerning the classification of nonlinear circuit elements and networks.

NOTE 2—The *N*-port $\boldsymbol{\mathcal{N}}$ is characterized by the collection of **all** possible admissible pseudo-wave vector pairs $\left(\mathbf{a}(t), \mathbf{b}(t)\right)$ under all possible forward pseudo-wave vector excitations $\mathbf{a}(t)$ for all $t_o \in \mathbf{R}$.

NOTE 3—The X-parameter® framework in its most general form is a special case of the admissible vector pair characterization of $\boldsymbol{\mathcal{N}}$. Specifically, the general X-parameter® represents the spectral mapping from all possible multi-tone pseudo-wave input spectrums (including dc, where it is assumed either voltage or current is the input). Which is a subset of all possible $\mathbf{a}(t)$ stimulus vector waveforms due to the finite number of tones—to one harmonic response (including dc, where we assume either voltage or current is the response) at a given port *n*. This is done so that by taking some finite set of these mappings to port *n* representing a set of harmonics, a time-domain response waveform $b_n(t)$ can be formed, which in turn, together with all of the remaining $N-1$ ports, results in the vector $\mathbf{b}(t)$ response. Assuming an equal set of K harmonics (plus dc) are used for each of the *N* ports, it follows that an admissible vector pair characterization constitutes $(K+1)N$ X-parameters®. In addition, the practical measurable set of X-parameters® used in commercial computer-aided engineering software and measurement instrumentation (usually denoted as $X^{(FB)}$, $X^{(S)}$, $X^{(T)}$, $X^{(FV)}$, $X^{(FI)}$, $X^{(Y)}$, and $X^{(Z)}$), which is a result of a spectral linearization of the general X-parameters® using a special subset of input stimuli at all the *N* ports, will thus also be a special case of the admissible vector pair characterization.

In 3.4.2 revise Equation (28) and Equation (29) as follows:

$$B_{p,k} = X_{p,k}^{(F)}\left(V_1,\left|A_{1,1}\right|,A_{2,1}P^{-1}\right)P^k + \sum_{q\geq 1,l>1} X_{p,k;q,l}^{(S)}\left(V_1,\left|A_{1,1}\right|,A_{2,1}P^{-1}\right)\cdot A_{q,l}P^{k-l}$$
$$+\sum_{q\geq 1,l>1} X_{p,k;q,l}^{(T)}\left(V_1,\left|A_{1,1}\right|,A_{2,1}P^{-1}\right)\cdot A_{q,l}^{*}P^{k+l}$$

$$I_p = X_p^{(I)}\left(V_1,\left|A_{1,1}\right|,A_{2,1}P^{-1}\right)+\sum_{q\geq 1,l>1} Re\left(X_{p;q,l}^{(Y)}\left(V_1,\left|A_{1,1}\right|,A_{2,1}P^{-1}\right)\cdot A_{q,l}\right) \tag{28}$$

$$B_{p,[n,m]} = X^{(F)}_{p,[n,m]}\left(V_1, \left|A_{1,[1,0]}\right|, \left|A_{2,[0,1]}\right|\right) P^{n}_{[1,0]} P^{m}_{[0,1]}$$
$$+ \sum_{q,n',m'} X^{(S)}_{p,[n,m];q,[n',m']}\left(V_1, \left|A_{1,[1,0]}\right|, \left|A_{2,[0,1]}\right|\right) P^{n-n'}_{[1,0]} P^{m-m'}_{[0,1]} A_{q,[n',m']}$$
$$+ \sum_{q,n',m'} X^{(T)}_{p,[n,m];q,[n',m']}\left(V_1, \left|A_{1,[1,0]}\right|, \left|A_{2,[0,1]}\right|\right) P^{n+n'}_{[1,0]} P^{m+m'}_{[0,1]} A^{*}_{q,[n',m']}$$

$$I_p = X^{(I)}_{p}\left(V_1, \left|A_{1,[1,0]}\right|, \left|A_{2,[0,1]}\right|\right) + \sum_{q,n',m'} Re\left(X^{(Y)}_{p;q,[n',m']}\left(V_1, \left|A_{1,[1,0]}\right|, \left|A_{2,[0,1]}\right|\right) \cdot P^{-n'}_{[1,0]} P^{-m'}_{[0,1]} A_{q,[n',m']} \right)$$

$$P_{[1,0]} = e^{j\phi\left(A_{1,[1,0]}\right)}$$

$$P_{[0,1]} = e^{j\phi\left(A_{2,[0,1]}\right)} \tag{29}$$

In 3.4.2 revise Equation (30), Equation (32), and the second paragraph after NOTE 3 as follows:

For example, consider the solution in the case of fundamental $A_{1,1}$ and $A_{2,1}$, and second harmonic $A_{2,2}$ large signal incoming-pseudo-wave phasors; in this case there are two relative phase terms $\left(Q_{2,1}/Q_{1,1}, Q_{2,2}/Q^{2}_{1,1}\right)$ and two magnitude terms $\left(\left|A_{2,1}\right|, \left|A_{2,2}\right|\right)$. The Cardiff Model thus requires four polynomial exponent terms $\left(m_{2,1}, n_{2,1}, m_{2,2}, n_{2,2}\right)$, along with the model coefficients, $K_{p,h,m_{2,1},n_{2,1},m_{2,2},n_{2,2}}$, of the respective polynomial expansion products, as shown in the Equation (30):

$$B_{p,h} = Q^{h}_{1,1}\left(\sum_{r_{2,2}=0}^{+\infty} \sum_{n_{2,2}=-\infty}^{+\infty} \sum_{r_{2,1}=0}^{+\infty} \sum_{n_{2,1}=-\infty}^{+\infty} K_{p,h,m_{2,1},n_{2,1},m_{2,2},n_{2,2}} \left|A_{2,1}\right|^{\left|n_{2,1}\right|+2r_{2,1}} \left(Q_{2,1}/Q_{1,1}\right)^{n_{2,1}} \left|A_{2,2}\right|^{\left|n_{2,2}\right|+2r_{2,2}} \left(Q_{2,2}/Q^{2}_{1,1}\right)^{n_{2,2}} \right) \tag{30}$$

$$B_{p,h} = Q^{h}_{1,1}\left(\sum_{r_{2,1}=0}^{\min\left(1,\left[(order-h)/2\right]\right)} \sum_{n_{2,1}=-\left(\left[(order-h)/2\right]-r_{2,1}\right)}^{h+\left(\left[(order-h)/2\right]-r_{2,1}\right)} K_{p,h,m_{2,1},n_{2,1}} \left|A_{2,1}\right|^{\left|n_{2,1}\right|+2r_{2,1}} \left(Q_{2,1}/Q_{1,1}\right)^{n_{2,1}} \right) \tag{32}$$