

Reflectionless Coupling Circuit for Broadband Power Line Communication

Gabriel N. de S. Ribeiro, André N. de S. Ribeiro, Luís G. da S. Costa, Carlos A. Duque, and Moisés V. Ribeiro

Abstract—This paper presents a reflectionless coupling circuit designed for broadband power line communication systems operating in low-voltage electric power grids. The circuit employs low-pass and high-pass pseudo-elliptic reflectionless filters to enhance impedance matching. The proposed circuit is evaluated against its non-reflectionless version using numerical simulations, prototype measurements, and on-grid test measurements. Results indicate that the proposed design maintains broadband transmission and significantly reduces reflection in both passband and stopbands. Additionally, measurement results show that the 100 Ω reference impedance design is more appropriate for electric power grid connections.

Keywords—Coupling circuit, impedance matching, power line communication, reflectionless filter.

I. INTRODUCTION

Power line communication (PLC) systems utilize existing electric power grids for data communication without requiring additional infrastructure [1]. However, data communication over electric power grids is challenging due to the dynamics of connected loads and equipment. These behaviors render dynamic access impedance, dynamic frequency-selective PLC channels, and high-power impulsive noise. As a result, the performance of PLC systems is often degraded, as extensively documented in the literature.

Due to the high-power nature of the mains frequency (i.e., 50/60 Hz), PLC transceivers require a coupling circuit to inject and extract data-carrying PLC signals. The coupling circuit blocks the mains frequency and provides electrical protection and electromagnetic isolation. Impedance matching across the entire frequency band is a relevant issue in coupling circuits, as improved impedance matching increases received signal power [2]. To achieve impedance matching, *a priori* studies have investigated high-pass and low-pass filters [3], transformer-based coupling structures with impedance adaptation [4], and resonant interfaces and broadband impedance-matching techniques [5]. Despite their achievements, PLC coupling circuits still rely on reflective circuit networks, in which out-of-band energy is returned to the source and impedance mismatching is not ensured. A promising research direction

to overcome these limitations is the adoption of reflectionless circuit networks. Reflectionless circuit networks enhance impedance matching and provide absorptive stopband behavior by dissipating rejected energy in internal terminations [6], potentially reducing standing-wave effects under frequency- and time-varying impedance conditions.

This study presents a reflectionless (RL) coupling circuit for broadband PLC systems operating in low-voltage electric power grids, targeting the 1.7-50 MHz frequency band as regulated by ANATEL in Brazil. The proposed RL PLC coupling circuit integrates low-pass and high-pass pseudo-elliptic RL filters to improve impedance matching. Numerical simulations demonstrate that the proposed RL PLC coupling circuit yield a small degradation in the signal transmission in comparison to its non-reflectionless (NRL) version. However, the proposal significantly reduces the magnitudes of the scattering parameters $|S_{11}|$ and $|S_{22}|$ compared to the NRL version, indicating superior absorptive characteristics and a substantial reduction in reflected power. Experimental results further confirm the enhanced performance of the RL PLC coupling circuit over the NRL version. Measurement results also indicate that a reference impedance of 100 Ω is more suitable for connecting the proposal to electric power grids.

II. REFLECTIONLESS FILTER

An RL filter can be derived from symmetry and circuit duality in even-/odd-mode decomposition of a symmetric two-port network [6]. Ideally, it yields a filter with zero reflection coefficient at all frequencies at both ports, while out-of-band energy is absorbed rather than reflected. Fig. 1(a) shows a symmetric two-port filter network, which can also be represented by a 2×2 scattering parameter matrix, while Fig. 1(b) shows the network bisected by its symmetry plane, producing two half-circuits. Under even-mode excitation, the two ports are driven with equal amplitudes and in phase; hence, no current crosses the symmetry plane, and the plane is modeled as a virtual open circuit, as shown in Fig. 1(c). Under odd-mode excitation, the two ports are driven with equal amplitudes and 180° phase difference; therefore, the symmetry plane is at zero potential and is modeled as a virtual ground, or virtual short circuit, as shown in Fig. 1(d). For a symmetric two-port reference impedance of Z_0 , the modal reflection coefficients are related to the two-port scattering parameters as follows [6]:

$$S_{11} = S_{22} = \frac{1}{2} (\Gamma_{\text{even}} + \Gamma_{\text{odd}}) \quad (1)$$

and

$$S_{21} = S_{12} = \frac{1}{2} (\Gamma_{\text{even}} - \Gamma_{\text{odd}}), \quad (2)$$

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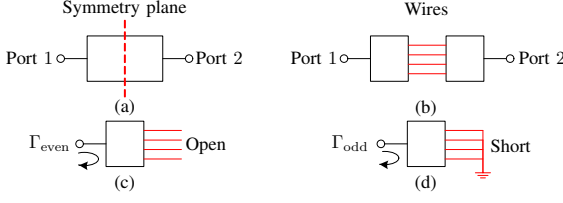


Fig. 1. Even/odd-mode decomposition of a symmetric two-port. (a) Symmetric two-port network, (b) shows the wires along the symmetry plane, (c) the even-mode equivalent circuit, and (d) the odd-mode equivalent circuit.

where Γ_{even} and Γ_{odd} are the reflection coefficients for even- and odd-mode equivalent circuits, respectively, as shown in Fig. 1(c) and Fig. 1(d). The reflectionless condition requires $S_{11} = S_{22} = 0$, and, consequently, results in $\Gamma_{\text{even}} = -\Gamma_{\text{odd}}$, which is obtained when the even- and odd-mode half-circuits are chosen as duals of one another. Substituting this condition into (2) yields $S_{21} = S_{12} = \Gamma_{\text{even}}$, meaning the transmission becomes fully determined by the even-mode response once matching holds [6]. Section III takes advantage of reflectionless filters to introduce a PLC coupling circuit for broadband PLC systems.

III. REFLECTIONLESS PLC COUPLING CIRCUIT

An appropriate design of a broadband PLC coupling circuit must ensure galvanic isolation between the electric power grid and the analog front-end (AFE) of the PLC transceiver, fast suppression of electrical transients from the electric power grid, analog filtering to define the frequency band, and low insertion loss and return loss. Usually, it is accomplished by a two-stage protection manner using gas metal oxide varistor (GMOV) and transient voltage suppressor (TVS) components, respectively, a blocking capacitor (C_{block}), a radio frequency (RF) transformer T_1 , and a combination of high-pass and low-pass analog filters.

Figure 2 presents the proposed RL PLC coupling circuit for a broadband PLC system. The novelty is the use of RL network structure is to implement a 5th-order low-pass and high-pass pseudo-elliptic RL filters¹. The pseudo-elliptic RL filter exhibits absorptive stopband behavior, dissipating out-of-band energy in internal resistive terminations rather than reflecting it to the source, and enhanced frequency selectivity, and, as will be shown in Section IV, increases tolerance to impedance variations when the coupling circuit operates under the dynamic impedance conditions of the electric power grid. Details about the design of pseudo-elliptic RL filters are presented in Subsection III-A.

A. Pseudo-Elliptic Reflectionless Filter

This subsection outlines the practical workflow for deriving a pseudo-elliptic RL filter² from a Chebyshev Type II RL one. In accordance with [6], [7] and considering the reference values (Z_0, ω_0), in which Z_0 is the reference impedance and ω_0 is the reference angular frequency, the lumped-element values are obtained. An elliptic-like approximation or pseudo-elliptic

RL filter can be accomplished by modifying the Chebyshev Type II RL by introducing a controlled passband ripple in the latter while retaining the RL condition [6]. Since a symmetric RL network satisfies $S_{21} = \Gamma_{\text{even}}$ when $S_{11} = 0$, the passband ripple requires the even-mode reflection magnitude to reach unity at a finite number of frequencies [6].

Figure 3 illustrates the main steps used to derive the pseudo-elliptic RL filter from the Chebyshev Type II one. The gray shading marks the cover-up region and the relocated elements. In Fig. 3(a), a series capacitor is inserted into the shunt-inductor branch of the even-mode circuit, forming a shunt series — LC branch. At its series-resonant frequency, this branch ideally exhibits zero impedance and creates the passband ripple maxima due to the RL condition. To preserve the RL property, the odd-mode network is then completed as the topological dual of the modified even-mode network [6]. Figures 3(b)–(e) realize this dual embedding through capacitor splitting, relocation into the cover-up region, symmetry-line insertion, and final dual completion. In Fig. 3(b), the first series capacitor is split, and part of it is relocated into the cover-up region. In Fig. 3(c), the symmetry line is inserted; in the odd-mode half-circuit, the added symmetry-line element appears as a shunt inductor, while the even-mode circuit remains unchanged. Figure 3(d) shows the splitting and partial relocation of the second series capacitor. Finally, Fig. 3(e) adds the second symmetry-line inductor, completing the odd-mode dual and yielding the pseudo-elliptic RL filter prototype [6].

The proposed RL PLC coupling circuit incorporates both high-pass and low-pass RL filters for the reference impedance of $Z_0 = 50 \Omega$, which is standard in RF designs. The procedure for determining the component values is as follows. The 5th-order pseudo-elliptic low-pass RL filter, with a cut-off frequency of 50 MHz and illustrated in Fig. 3(e), was designed based on the prototype low-pass RL filter with a cut-off frequency of 1 GHz presented in [6, p. 238]. The component values were calculated by applying a component value-scaling factor of 20 to shift the cut-off frequency from 1 GHz to 50 MHz. The 5th-order pseudo-elliptic high-pass RL filter, with a cut-off frequency of 1.7 MHz, was designed by first converting the prototype in [6, p. 238] into a pseudo-elliptic high-pass RL filter with a cut-off frequency of 1 GHz, followed by the application of a frequency scaling factor of 588.24 to the component values. The resulting component values and their commercial equivalents used in the implementation are in Table I. Information on the other components of the proposed circuit is also included in this table.

Now, if the electric power grids use two-conductor power cables with a cross-sectional area of 2.5 mm^2 , $D \approx 4 \text{ mm}$ being the center-to-center distance between the conductors, and $a = \sqrt{2.5/\pi} \text{ mm}$ being the conductor radius, then the characteristic impedance is given by

$$Z_{c,0} = \frac{120}{\sqrt{\epsilon_r}} \operatorname{arccosh}\left(\frac{D}{2a}\right) \approx 100 \Omega, \quad (3)$$

where $\epsilon_r \approx 3$ is the relative permittivity of the insulating material. To comply for $Z_0 = 100 \Omega$ without redesigning the proposal, the transformer T_1 , see Fig. 2, with impedance ratios equal to 1:1 is replaced by another with 1:2 ratio.

¹According to [7], pseudo-elliptic RL filters are simpler to obtain.

²A true elliptic transfer function does not admit an exact RL realization.

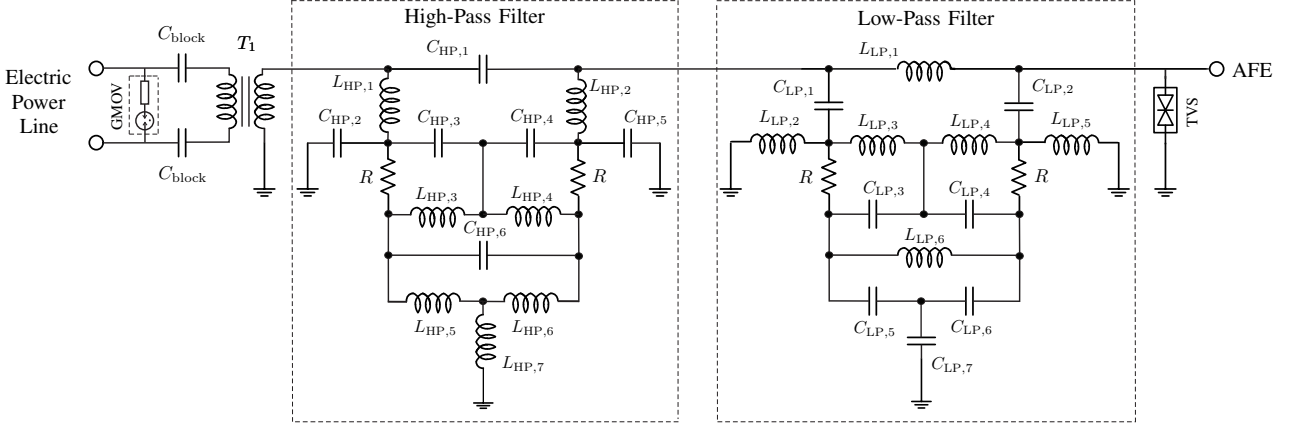


Fig. 2. The schematic of the proposed reflectionless and capacitive PLC coupling circuit for broadband applications.

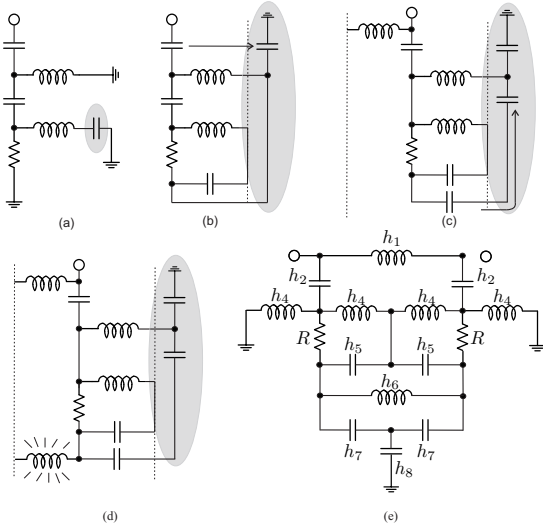


Fig. 3. Main steps for deriving a pseudo-elliptic reflectionless filter: (a) adding a series capacitor to the shunt-inductor branch; (b) splitting the first series capacitor; (c) inserting the symmetry line; (d) splitting the second series capacitor; and (e) adding the second shunt inductor.

IV. NUMERICAL ANALYSIS

This section presents numerical simulation results in terms of scattering-parameter magnitudes ($|S_{11}|$, $|S_{22}|$, $|S_{12}|$, $|S_{21}|$) for the proposed RL PLC coupling circuit and its NRL version, both analyzed with the Advanced Design System (ADS) software from Keysight. The sole distinction between the RL and NRL PLC coupling circuits is that the former incorporates the RL filters described in Subsection III-A, whereas the latter employs 5th-order pseudo-elliptic high-pass and low-pass filters with cut-off frequencies of 1.7 MHz and 50 MHz, respectively, designed in ADS for $Z_0 = 50 \Omega$. All numerical simulations utilized commercially available component values.

For $Z_0 = 50 \Omega$ as the reference impedance at both ports, Fig. 4 displays a comparison of the scattering-parameter magnitudes for the proposed RL and NRL PLC coupling circuits. Both circuits exhibit $|S_{21}| = |S_{12}|$ values close to 0 dB, yet their impedance matching characteristics differ markedly. In the lower stopband, below 1.7 MHz, the NRL circuit remains highly reflective, with $|S_{11}| \approx 0$ dB, while the

TABLE I
CALCULATED AND ADOPTED COMMERCIAL VALUES OF THE
PSEUDO-ELLIPTIC RL FILTERS FOR $Z_0 = 50 \Omega$.

HIGH-PASS 1.7 MHz			LOW-PASS 50 MHz		
Component	Calculated value	Commercial value	Component	Calculated value	Commercial value
$C_{HP,1}$	1.33 nF	1.5 nF	$L_{LP,1}$	224 nH	220 nH
$L_{HP,1}$	6.7 μ H	8.2 μ H	$C_{LP,1}$	44 pF	47 pF
$L_{HP,2}$	6.7 μ H	8.2 μ H	$C_{LP,2}$	44 pF	47 pF
$C_{HP,2}$	2.65 nF	3.9 nF	$L_{LP,2}$	112 nH	110 nH
$C_{HP,3}$	286 pF	270 pF	$L_{LP,3}$	1.04 μ H	1 μ H
$C_{HP,4}$	286 pF	270 pF	$L_{LP,4}$	1.04 μ H	1 μ H
$C_{HP,5}$	2.65 nF	3.9 nF	$L_{LP,5}$	112 nH	110 nH
R	50 Ω	50 Ω	R	50 Ω	50 Ω
$L_{HP,3}$	29.7 μ H	33 μ H	$C_{LP,3}$	10 pF	10 pF
$L_{HP,4}$	29.7 μ H	33 μ H	$C_{LP,4}$	10 pF	10 pF
$C_{HP,6}$	5.95 nF	6.8 nF	$L_{LP,6}$	50 nH	47 nH
$L_{HP,5}$	747 nH	750 nH	$C_{LP,5}$	400 pF	300 pF
$L_{HP,6}$	747 nH	750 nH	$C_{LP,6}$	400 pF	300 pF
$L_{HP,7}$	3.31 μ H	3.3 μ H	$C_{LP,7}$	90 pF	82 pF

$$C_{\text{block}} = 2.20 \mu\text{F}$$

$$T_1 \text{ for } Z_0 = 50 \Omega: \text{PWB-1010-1L}$$

$$T_1 \text{ for } Z_0 = 100 \Omega: \text{WB2-1-2WL}$$

proposed RL PLC coupling circuit reduces input reflection to approximately -35 to -40 dB. Near 1.7 MHz, both circuits display the anticipated transition; however, the proposed design sustains significantly lower $|S_{11}|$ and $|S_{22}|$ values. Within the passband, the NRL circuit exhibits considerable $|S_{22}|$ fluctuations between -10 and -40 dB, whereas the proposed circuit achieves a more uniform absorptive response, with $|S_{11}|$ typically ranging from -25 to -35 dB. Above 50 MHz, both circuits enter the upper stopband and $|S_{21}| = |S_{12}|$ decreases, but the proposed design continues to maintain lower $|S_{11}|$ and $|S_{22}|$ up to 100 MHz. Overall, numerical simulations show that the RL PLC coupling circuit supports broadband transmission and enhances impedance matching, particularly outside the 1.7–50 MHz frequency band.

V. EXPERIMENTAL ANALYSIS

The performance of the proposed RL coupling circuit prototype and its NRL version is evaluated by measuring

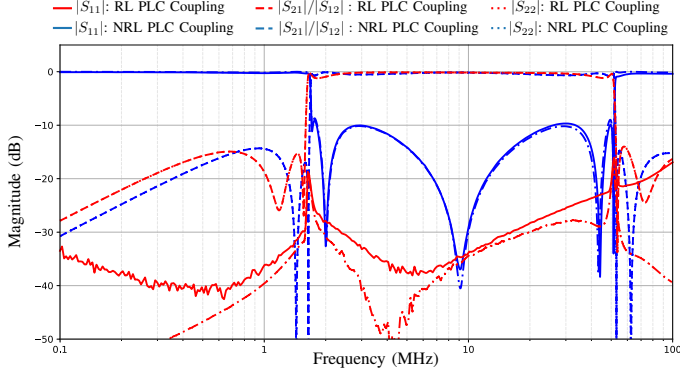


Fig. 4. Comparison of scattering parameter magnitudes between the proposed RL PLC coupling circuit and the NRL PLC coupling circuit.

$|S_{11}|$, $|S_{22}|$, $|S_{12}|$, and $|S_{21}|$ with a Rohde & Schwarz FSH-8 vector network analyzer (VNA). Figure 5 shows that both prototypes were assembled on a double-sided FR-4 printed circuit board (PCB).

Figure 6 shows the measured scattering parameters for both prototypes. Within the transmission band (1.7–50 MHz), both prototypes yield low insertion loss, with $|S_{21}|$ and $|S_{12}|$ values near -0.8 dB. Below 1.7 MHz, the prototype of the NRL PLC coupling circuit exhibits strong mismatch, as indicated by $|S_{11}|$ and $|S_{22}|$ values close to 0 dB. In contrast, the prototype of the proposed RL PLC coupling circuit achieves significantly lower reflected power: $|S_{11}|$ ranges from -18 to -26 dB between 0.1 and 1 MHz, and $|S_{22}|$ remains at or below -20 dB in the same stopband region. At the transition band, $|S_{11}|$ increases, yet the RL circuit prototype maintains superior matching. Throughout the passband, the proposed RL PLC coupling circuit prototype sustains input and output reflections near -20 dB, whereas the NRL prototype displays broader reflection variations, with $|S_{11}|$ ranging from -10 to -20 dB. Above 50 MHz, both prototypes enter the upper stopband and exhibit ripple; however, the proposed RL PLC coupling circuit prototype predominantly remains below -15 dB for $|S_{11}|$ and $|S_{22}|$, while the NRL prototype stays near 0 dB. These measured results confirm that the proposed RL PLC coupling circuit achieves improved broadband matching, both within and outside the transmission band, without significant transmission degradation.

Fig. 7 illustrates the experimental setup used to measure the scattering-parameters of both PLC coupling circuits connected to the electric power grid under on-grid conditions. In this arrangement, the line impedance stabilization network (LISN) decouples the measurement setup from disturbances present on the electric power grid and provides a suitable interface for connecting test instruments, such as a VNA, to a two-branch power line, with each branch having a length of $\ell = 10$ m.

Fig. 8 shows the attained results with the proposed RL PLC coupling circuit and its NRL version by considering the two transformer-based matching conditions discussed in Section III-A, to comply with the reference impedance of $Z_0 = 50 \Omega$ and $Z_0 = 100 \Omega$ when the experimental setup is used.

The measured $|S_{11}|$ shown in Fig. 8(a) highlights the input-

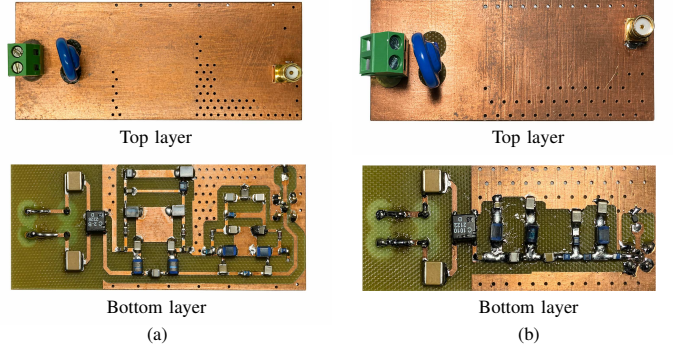


Fig. 5. Prototypes of (a) the RL PLC coupling circuit and (b) the NRL PLC coupling circuit.

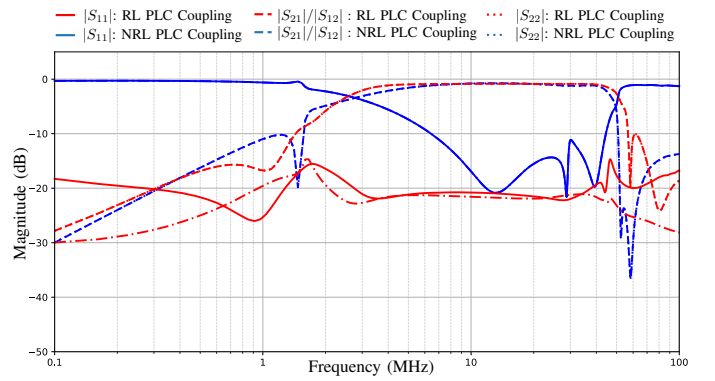


Fig. 6. Comparison of scattering parameter magnitudes for the prototypes of RL and NRL PLC coupling circuits.

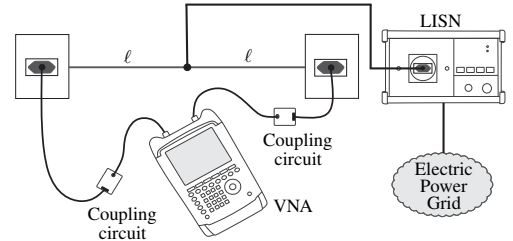


Fig. 7. An illustration of the experimental setup to measure prototypes of the RL and NRL PLC coupling circuits.

matching advantage of the proposed RL PLC coupling circuit over its NRL version in the three measured frequency regions. In the lower stopband (0.1–1.7 MHz), $|S_{11}|$ for the NRL version remains close to 0 dB for both 50Ω and 100Ω , while the proposal yields $|S_{11}|$ and $|S_{22}|$ between -22 dB and -18 dB, respectively. In the passband (1.7–50 MHz), the NRL version exhibits more ripple and resonant nulls, whereas the proposal attains a smoother response, typically between -8 and -15 dB for 100Ω and between -10 and -20 dB for 50Ω . In the upper stopband (50–100 MHz), $|S_{11}|$ for the NRL version returns to 0 dB, while the proposal remains better matched, with $|S_{11}|$ roughly between -15 and -20 dB for 100Ω and around -15 dB for 50Ω . Overall, changing the impedance reference from 50Ω to 100Ω preserves the useful PLC band, but clearly modifies the reflected power at both

ports.

Fig. 8(b) shows that the measured $|S_{22}|$ follows the same behavior of $|S_{11}|$. From 0.1 to 1.7 MHz, $|S_{22}|$ of the NRL version is close to 0 dB for both 50 Ω and 100 Ω , whereas the proposed RL PLC coupling circuit yields $|S_{22}|$ between -15 and -20 dB. In the passband (1.7–50 MHz), the proposal generally keeps $|S_{22}|$ between -10 and -20 dB for 100 Ω and around -5 to -15 dB for 50 Ω , while the NRL version shows higher reflection and resonant variations. From 50 to 100 MHz, the NRL version again approaches 0 dB, whereas the proposed RL PLC coupling circuit maintains better output reflection, especially for the 100 Ω case.

The measured S_{21} and S_{12} magnitudes shown in Fig. 8(c) confirm the proposed RL PLC coupling circuit and its NRL version preserve the broadband transmission in the passband (1.7–50 MHz) for both reference impedances. From 0.1 to 1.7 MHz, the broadband transmission remains substantially attenuated for both 50 Ω and 100 Ω , with an increase near the transition band. In the passband, the transmission is typically between -5 and -10 dB for both reference impedances; the 100 Ω curves show slightly better mid-band flatness, whereas the 50 Ω cases present more ripple and localized resonant variations. From 50 to 100 MHz, $|S_{21}|$ and $|S_{12}|$ decrease again, confirming the upper-band rejection, with no severe degradation introduced by the reference impedance of 100 Ω .

Overall, the proposed RL PLC coupling circuit achieves a better balance between transmission and matching across passband and stopbands. Also, it remarkably outperforms its NRL version. In the stopband ranges (0.1–1.7 MHz and 50–100 MHz), it substantially reduces reflections while incurring a small decrease in the stopband attenuation. In the passband (1.7–50 MHz), reflections is remarkably reduced and transmission shows a small degradation in comparison to the NRL version. Also, a reference impedance of 100 Ω generally improves input and output matching by reducing $|S_{11}|$ and $|S_{22}|$ compared to 50 Ω , with negligible impact on $|S_{21}|$ and $|S_{12}|$. Therefore, a 100 Ω reference impedance, achieved with a 1:2 transformer ratio, is more suitable when the coupling device is connected to the electric power grid. Conversely, a 50 Ω reference impedance is consistent with RF standard design practices.

VI. CONCLUSION

Aiming to reduce energy reflections in connections of PLC transceivers to electric power grids, a reflectionless coupling circuit for PLC system operation in the frequency band of 1.7–50 MHz was introduced. Numerical simulations, prototype measurement results, and on-grid measurements showed that the proposed RL PLC coupling circuit outperforms its NRL counterpart in terms of scattering parameter magnitudes across both the passband and stopbands. Furthermore, on-grid measurements indicate that a design based on a reference impedance of 100 Ω is more suitable for electric power grid connections.

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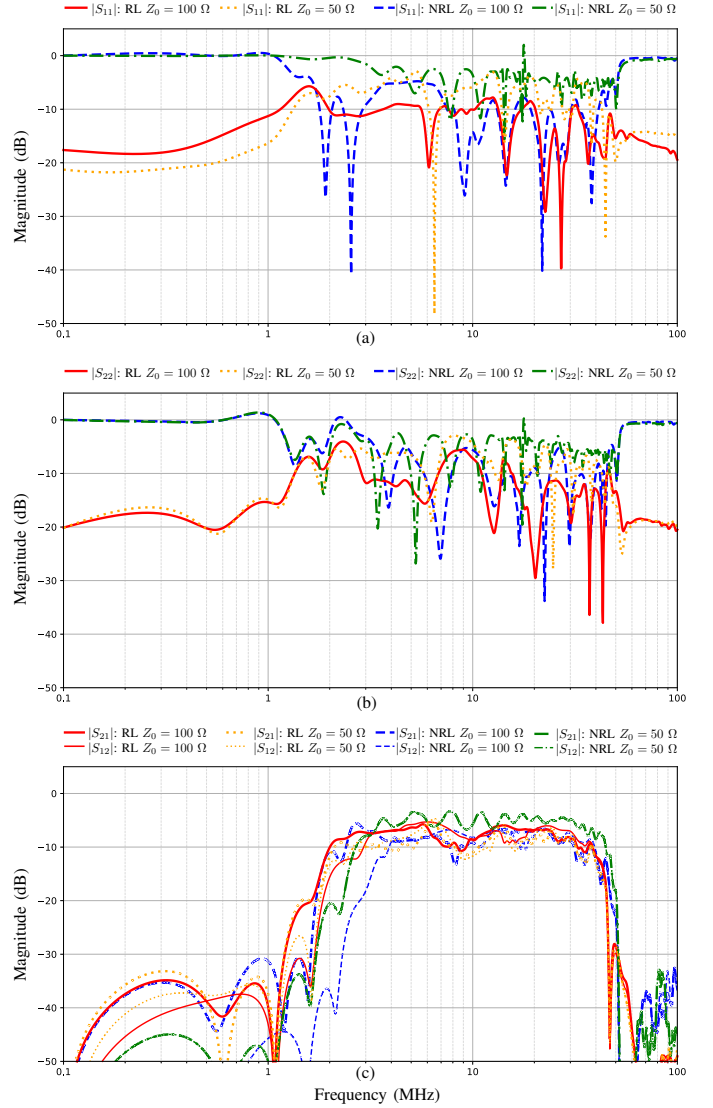


Fig. 8. Comparison of scattering-parameter magnitudes of prototypes for RL PLC and NRL PLC coupling circuits, using reference impedances of 50 Ω and 100 Ω .

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