gain of 62.5 dBΩ (50 Ω loaded at each output, photoreceiver capacitance of 0.15 pF), a −3 dB cutoff frequency of 5.6 GHz, and an input impedance of 150 Ω, shown in Figure 3.

Moreover, an input-equivalent noise current of 12.3 pA/√Hz in the 100 MHz–7.5 GHz band has been evaluated by simulation, using the foundry library noise models for active and passive devices.

5. CONCLUSION
A novel topology of single-input to differential-output converters suitable for designing differential output transimpedance amplifiers has been proposed.

The converter provides 6 dB extra gain with respect to a simple differential pair, and allows us to design transimpedance amplifiers featuring improved performance in terms of conversion gain, input impedance, and offset compensation with respect to commonly used transimpedance topologies.

A front-end amplifier for SDH systems operating at 10 Gbit/s has been designed and fabricated in Philips PML ED02AH GaAs HEMT technology (0.2 μm gate length).

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REFERENCES

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the true frequency response of a wire antenna. One example that was considered is a ½ m long center-fed wire antenna with a 1 mm radius. In this case, the fidelity of the equivalent circuit presented in [2] was very poor, and consequently, some form of circuit optimization is required.

A genetic-algorithm technique was then used to optimize the equivalent circuit of the dipole antenna. The GA views the target antenna impedance, computed from a method-of-moments (MoM) computer code [4], as the design objective, and strives to select component values of the equivalent circuit to best match the desired impedance curves. A 10 bit binary-coded gene represents the value of each of the components. Therefore, a candidate antenna equivalent circuit is represented by a single 160 bit chromosome. The population size in this example is chosen to be equal to the chromosome size. The objective function is given by

\[
F = \sum_{n=1}^{N_f} \left[ \left( \text{Re}^{\text{MoM}}(f_n) - \text{Re}^{\text{GA}}(f_n) \right)^2 + \left( \text{Im}^{\text{MoM}}(f_n) - \text{Im}^{\text{GA}}(f_n) \right)^2 \right]
\]

where \( N_f \) is the number of frequency points, and \( \text{Re}^{\text{MoM}}(f_n) \) and \( \text{Im}^{\text{MoM}}(f_n) \) are the real and imaginary parts of the antenna impedance computed by the method of moments. \( \text{Re}^{\text{GA}}(f_n) \) and \( \text{Im}^{\text{GA}}(f_n) \) are the real and imaginary components of the equivalent circuit model chosen by the GA.

Figure 2 depicts the optimized results, which clearly illustrate a good agreement between the impedance of the genetically optimized antenna equivalent circuit and the antenna impedance computed by MoM. The model component values selected by the GA are listed below:

- \( R_{11} = 3.92 \, \text{k}\Omega \)
- \( R_{12} = 0.903 \, \Omega \)
- \( C_{11} = 1.21 \, \text{pF} \)
- \( L_{11} = 261 \, \text{nH} \)
- \( R_{21} = 5.36 \, \text{k}\Omega \)
- \( C_{21} = 0.241 \, \text{pF} \)
- \( L_{21} = 206 \, \text{nH} \)
- \( R_{31} = 4.24 \, \text{k}\Omega \)
- \( C_{31} = 0.183 \, \text{pF} \)
- \( C_{32} = 0.100 \, \text{pF} \)
- \( L_{31} = 186 \, \text{nH} \)
- \( R_{41} = 1.24 \, \text{k}\Omega \)
- \( C_{41} = 0.185 \, \text{pF} \)
- \( C_{42} = 0.367 \, \text{pF} \)
- \( L_{41} = 35.4 \, \text{nH} \)

One disadvantage of the lumped-element model is the fact that each additional overtone resonance requires another circuit branch and the recalculation of all model elements. Furthermore, the sensitivity of the model component values seems to increase with the addition of overtone circuit branches. This is a fundamental problem rooted in the fact that antennas have overtone responses, but lumped circuit elements do not.

### III. GA OPTIMIZATION OF LUMPED-COMPONENT TRANSMISSION-LINE EQUIVALENT MODEL

It then makes sense to consider an antenna equivalent-circuit model built around components that, like the target antenna, have periodic frequency-domain impedance behavior. Figure 3 shows such a model consisting of a pair of transmission-line sections—one mismatched into a high-impedance load, the other mismatched into a low-impedance load—in series with a capacitor. This simple model has repetitive reactance poles and zeros very similar to an actual dipole. The series capacitor improves the model accuracy at low frequencies. Both transmission lines are about one-eighth wavelength long at the first half-wave resonance.
To further improve the model fidelity, the lumped-impedance loads are added at both ends of the transmission lines. Figure 4 shows an antenna equivalent-circuit model with improved impedance fidelity where negative capacitors are included in the load networks. Because negative capacitors have positive reactance, they shift the full-wave resonance frequencies downward, and because the magnitude of this reactance decreases with frequency, the fundamental resonance is affected more strongly than the overtones, thereby improving the correspondence between the circuit model and actual antenna impedance poles. The transmission lines themselves are allowed to be lossy, with the loss increasing with frequency. Again, the GA is applied to select the parameter values needed for model accuracy. The transmission-line segments are modeled using the standard hyperbolic trigonometric relationship with a complex propagation constant

\[ \gamma = \alpha + j\beta, \quad \beta = 2\pi/\lambda, \quad \alpha \text{ is one of the parameters which will be optimized by the GA.} \]

Figure 5 shows the optimized results. The component values selected by the GA are given below:

- \( R_{11} = 13.11 \Omega \)
- \( R_{12} = 3600 \Omega \)
- \( R_{13} = 500 \Omega \)
- \( C_1 = 2.504 \text{ pF} \)
- \( C_2 = -16.25 \text{ pF} \)
- \( C_3 = 0.4000 \text{ pF} \)
- \( C_4 = 0.1388 \text{ pF} \)

\[ \text{line } L_1: \quad Z_n = 214.8 \Omega, \quad \text{length } = 0.1248 \text{ M}, \]
\[ \text{line } L_2: \quad Z_n = 195.1 \Omega, \quad \text{length } = 0.1304 \text{ M}, \]
\[ \alpha = 0.0744 + 3.000 \log_{10} (f/900 \text{ MHz}) \]
\[ \alpha = 0.0101 + 0.0339 \log_{10} (f/90 \text{ MHz}) \]

We noticed that transmission line \( L_1 \) dominates the odd-order full-wave resonances. Altering \( L_1 \) or its associated load networks changes the first and third full-wave resonances with little effect on the second and fourth. The second line and associated lumped components chiefly affect the even-order resonances. Series RC branches have an increasing effect at higher frequencies.

IV. CONCLUSIONS

We have demonstrated the use of a genetic algorithm to improve the impedance fidelity of two broadband antenna equivalent-circuit models for a dipole. In the first example, a GA is used to optimize component values in a lumped RLC network. In the second example, the authors present a novel transmission-line-based antenna equivalent-circuit model, and again determine optimum model parameters with the use of a genetic algorithm. The second equivalent-circuit model exhibits quite good impedance fidelity over a bandwidth exceeding five octaves, including the fundamental through the fourth overtone response.

REFERENCES


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