

# An Ultra-Wideband Microwave Balun using a Tapered Coaxial Coil Structure working from kHz range to beyond 26.5 GHz.

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**Abstract** — Many wideband baluns have been presented in the past using coupled lines, pure magnetic coupling or slotlines. Their limitations were set whether in high frequency or low frequency performance. Due to their lumped element bandpass representation many of them allow just certain bandwidth. The tapered coaxial coil structure allows balun operation beyond 26 GHz and down to the kHz range through partial ferrite filling. The cable losses, cable cut-off frequency, the number of windings, the permeability of the ferrite and the minimum coil diameter limit the bandwidth. The tapering allows resonance free operation through the whole band. Many microwave devices like mixers, power amplifiers, SWR-bridges, antennas, etc. can be made more broadband with this kind of balun. A stepwise approach to the proposed structure is presented and compared to previous balun implementations. Finally a measurement is provided and some implementation possibilities are discussed.

**Index Terms** — broadband, balun, coaxial cable.

## I. INTRODUCTION

The idea of coaxial baluns has been used so far mainly in broadband power amplifiers especially in push-pull designs for input and output coupling. Whether there is a ferrite sleeve covering a straight piece of coaxial cable or it is wound around a toroidal ferrite core [3]-[5]. While the ferrite material ensures low frequency operation, the upper frequency limit is determined by the physical construction of the cable assembly due to resonances. To understand the frequency behavior of balun constructions, theoretical fundamentals of operation have to be found starting with a simple coaxial line embedded in a metal cylinder with a quarter wavelength in length. The principal setup is shown in Figure 1. A coaxial line with a characteristic impedance of  $Z_1$  is embedded in a hollow metal cylinder maintaining a characteristic impedance  $Z_0$  between the outer coaxial conductor and the surrounding metal. For calculation simplicity it is supposed that there is no material involved with a relative permittivity greater than one. R. J. Wenzel [1] showed an efficient way for calculating the Z-Matrix of coupled line structures by determining it's static capacitance equivalent circuit.

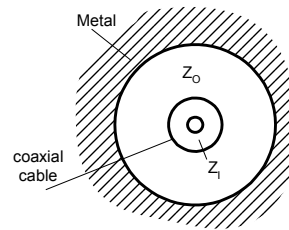


Fig. 1. Coaxial line with a characteristic impedance  $Z_1$  embedded in a hollow metal cylinder with a characteristic impedance  $Z_0$  between the outer coaxial conductor and the surrounding metal

It is based on the fact that the characteristic impedance of a transmission-line is directly related to its line capacitance per unit length, which is defined as:

$$Z_0 = \frac{\eta_0}{\sqrt{\epsilon_r} \frac{C'}{\epsilon}} = \frac{\eta_0}{\sqrt{\epsilon_r} C} \quad (1)$$

While  $C$  specifies the dimensionless static self- and mutual-capacitance per unit length, which exists between the open-circuited coupled, lines and ground.  $\eta_0$  stands for the free space characteristic impedance. Figure 2 now documents the static capacitance substitution of the physical construction out of Figure 1.

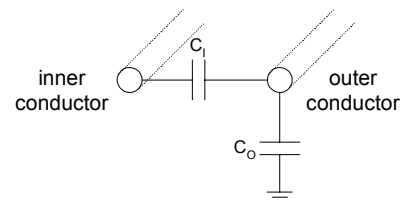


Fig. 2. Capacitance substitution of the circuit in Figure 1.

The left circle in Figure 2 represents the inner conductor, while the right circle stands for the outer conductor of the coaxial cable. As all of the electric field lines of the inner conductor end on the outer conductor, there is no capacitance between the inner conductor and ground. This fact also explains the reason why the coupling coefficient  $K$  between inner and outer coaxial conductor is 1.

The Z-Matrix of the coupled structure out of Figure 1 is presented in formula (2) below

$$\frac{1}{S} \begin{pmatrix} Z_1 + Z_o & Z_o & (Z_1 + Z_o)\sqrt{1-S^2} & Z_o\sqrt{1-S^2} \\ Z_o & Z_o & Z_o\sqrt{1-S^2} & Z_o\sqrt{1-S^2} \\ (Z_1 + Z_o)\sqrt{1-S^2} & Z_o\sqrt{1-S^2} & Z_1 + Z_o & Z_o \\ Z_o\sqrt{1-S^2} & Z_o\sqrt{1-S^2} & Z_o & Z_o \end{pmatrix}$$

The term S represents the Richard transformed frequency variable:

$$S = j \tan\left(\frac{\pi \cdot f}{f_0}\right) \quad (3)$$

Port 1 in the Z Matrix representation (3) is conjunct with the inner conductor at the physical position  $z = 0$ , Port 2 represents the outer conductor at  $z = 0$ . Port 3 and Port 4 are the inner and outer conductor terminals at the position  $z = \lambda_0/4$  respectively. The wavelength  $\lambda_0$  is described by the frequency  $f_0$  in equation (3) as:

$$\frac{\lambda_0}{4} = \frac{c}{f_0} \quad (4)$$

For correct balun operation the coaxial structure needs to be connected and wired as shown in Figure 3.

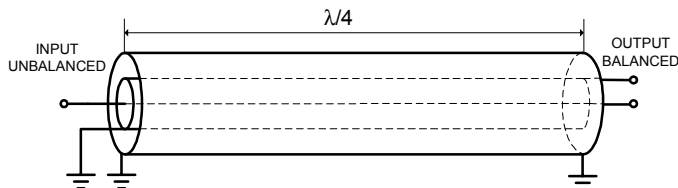


Fig.3. Interconnection of the coaxial-structure for balun operation.

It is assumed now that the characteristic impedance  $Z_1$  of the coaxial cable is 50 Ohms, and the frequency  $f_0$  is 1 GHz, the source impedance is also 50 Ohms, and the two termination impedances of the balun are 25 Ohms each. A frequency plot versus insertion loss can be calculated with these values. Figure 4 shows the frequency behavior of the balun with two different  $Z_o$  values. The two curves with bigger line width in Figure 4 represent balun operation with an outer characteristic impedance of 50 Ohms. The upper trace refers to the inner coaxial conductor and the lower trace to the outer coaxial conductor at the balun output respectively. It can be clearly seen that good balun performance is only possible between 800 MHz and 1200 MHz. If the characteristic impedance between the outer coaxial conductor and the surrounding metal is increased

to  $Z_o = 400$  Ohms which is shown in Figure 4 by the two traces with smaller width, a much broader frequency performance is achieved.

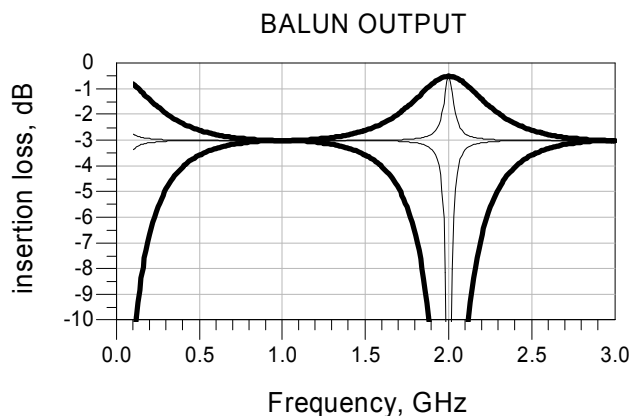


Fig. 4. Balun frequency response with  $Z_o = 50$  Ohms (thick lines) and  $Z_o = 400$  Ohms (thin lines)

With this high  $Z_o$  value the balun is working from 300 MHz to 1700 MHz. Increasing  $Z_o$  to improve balun performance does also mean to physically increase the distance between the outer coaxial conductor and the surrounding metal. As this stands in hard contrast with a small physical size and practical realization, another coaxial structure must be found to satisfy the condition  $Z_o \gg Z_1$ .

## II. COAXIAL COIL STRUCTURE

A coil of wire has a higher inductance than a straight wire of the same length, through magnetic coupling effects between the windings. Electrical coupling between the windings and ground lead to a lower "self resonance" frequency than the straight wire at it's resonance at  $\lambda_0/4$ . This configuration is especially advantageous for the coaxial balun. If the coaxial cable is wound like an inductor the outer characteristic impedance increases, which leads to an improved bandwidth characteristic, while the size can be kept small. To analyze such coil structures a fitting theory has to be found. A general transmission-line theory can be applied if the "N-windings-coil"-problem can be transformed to finding the setup of a Z-Matrix of N-parallel coupled lines. To simplify the theory for now instead of a coaxial cable, a simple wire is used. Figure 5a shows a coil with three windings. This coil can be transformed to an equivalent model with parallel-coupled open wire-rings, which is shown in Figure 5b. Ideal wires to simulate a coil structure and so to keep the direction of current-flow constant

connect the rings. This structure can be flattened to a simple coupled line model as shown in figure 5b without introducing too many errors. The physical length of each line in Figure 5b needs to be the same as the circumference of one coil-ring in Figure 5a.

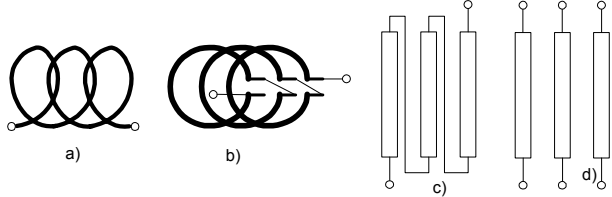


Fig.5. a) Three winding wire coil b) Interconnected parallel wirings to form a coil-structure electrically equivalent to case a, c) Equivalent plane coupled line structure to substitute case b, d) Plane coupled line structure as described by the static capacitor network.

Like already mentioned above and derived by [1], a static capacitor network can represent the coupled line structure. If these simple coupled transmission-lines are replaced now by coupled coaxial lines, a configuration as shown in Figure 6 appears. Three coaxial-coupled lines are connected together with the output of the first cable joined to the input of the second cable and the output of the second cable connected to the input of the third cable. Some additional parameters are added to the simple configuration of Figure 1: The mutual characteristic impedance between two lines  $Z_c$  is introduced and represents the coupling capacitor  $C_c$  through formula (1), the second variable is the cable length per winding and at

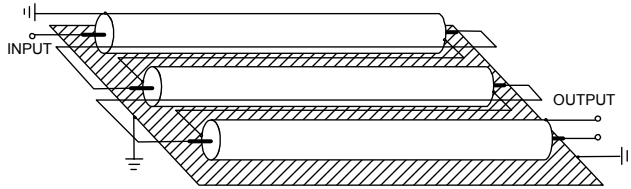


Fig.6. Transformed coupled coaxial line balun.

last the number of coupled structures. Knowing the structure in Figure 6, a substitute capacitor network can be found. Figure 7 shows a universal static capacitor network for an arbitrary number of open circuited-coupled coaxial lines. The  $Z$ -parameter-matrix of a coaxial coil balun with eight turns ( $N = 8$ ) and  $Z_o = 200$  Ohms,  $Z_1 = 50$  Ohms,  $Z_c = 35$  Ohms, winding length = 9.375 mm (nominal  $f_0 = 1$  GHz) will have an order of 32, and has been derived from the network out of Figure 7. A linear  $Z$ -parameter based solver algorithm for arbitrary matrix sizes [2] has been used to calculate the frequency behavior of the balun. Figure 8 shows the output of the simulation.

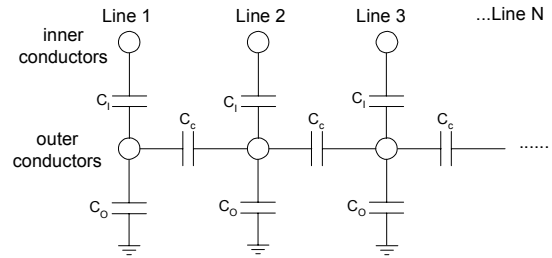


Fig.7. Static capacitance network for an arbitrary number of coupled coaxial lines (not connected as coil structure).

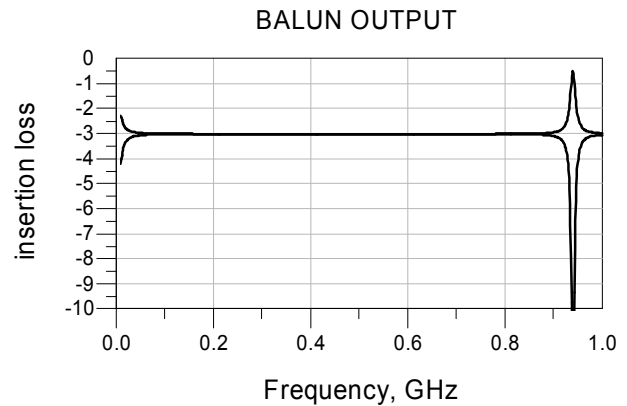


Fig.8. Frequency behavior of the coaxial coil balun.

The result shows that the relative bandwidth has been increased from 141 % ( 300 MHz to 1700 MHz with  $Z_o = 400$  Ohms) to 185 % (30 MHz to 900 MHz working range with  $Z_o = 200$  Ohms). This is equivalent to a single line balun like in Figure 1 with  $Z_o = 1000$  Ohms and  $f_0 = 0.47$  GHz. This proves that the coil structure increases the relative balun bandwidth. However resonance free operation has not been achieved yet.

### III. TAPERING THE COAXIAL COIL

If the coaxial coil is tapered, so that its shape forms a cone, each winding has a different diameter and so wire-length. That for the coupling between the lines can not be calculated through simple coupled transmission-line theory. Figure 9a shows the structure of a tapered coaxial coil. The outer conductor of the coaxial cable must be insulated to prevent electrical conduction between the windings. Through the different winding diameters no resonant structure is formed and the resonant peaks shown in Figure 8 are the better suppressed the higher the coupling between the windings become.

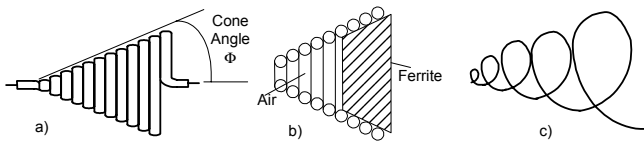


Fig.9. a) Physical structure of the tapered coaxial coil balun b) Partial ferrite filling for low frequency operation c) winding scheme of the tapered coaxial balun.

The smallest winding diameter determines the upper frequency limit of balun operation, the lower frequency is set by the number of turns and so by low frequency inductance. By partial ferrite filling as shown in Figure 9b, the lower frequency limit can be shifted to the kHz range. The ferrite's position ensures that losses through the material's high  $\mu''$  do not occur, as for reasonable high frequencies the air "inductor" in front of the ferrite filled part produces enough  $|Z|$  so that the balun transformation is fully done in this section. A cone angle of  $15^\circ$  to  $20^\circ$  showed best performance. A balun has been manufactured using UT-12, 0.35-mm diameter coaxial cable with 30 turns and a cone angle of about  $20^\circ$ . The size of this balun is approximately 13 mm length and a maximum diameter of 6 mm. The performance has been measured through a two coil setup as shown in Figure 10 using a HP 8510 network analyzer.

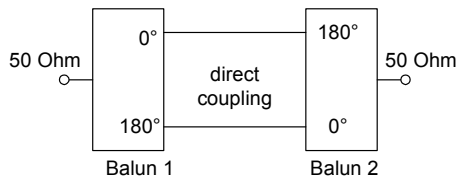


Fig.10. Two balun configuration for insertion loss and performance measurement.

#### IV. RESULTS

Figure 11 shows the measurement result of the coaxial coil structure with a configuration shown in Figure 10. The maximum insertion loss of about 8 dB at 26.5 GHz is mainly caused by the cable losses and the setup. It can be assumed that a single balun has a maximum loss of about 4 dB. The balun can be used for single- and double-balanced ultra-broadband mixers, wideband power amplifiers, SWR-bridges or wideband antennas. A thicker cable up to RG-402 can be used for power-amplifier splitters and combiners, but the upper useable frequency

limit will be decreased down to 10 GHz to 15 GHz. The number of windings can be arbitrarily increased; it is even possible to wind the cable subsequent to the conical coaxial coil around a toroidal ferrite core to lower the useable frequency into the Hz range. The balanced port is always designated to the sharp side of the cone. To get the best performance it is necessary to have ground as far as possible away from the coil.

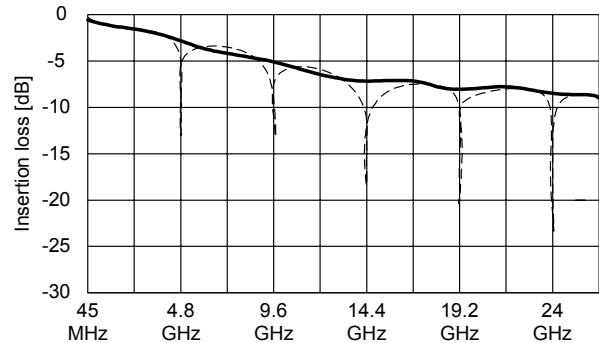


Fig.11. Thick solid line: Resonance free insertion loss of the double balun configuration of Figure 11 using the tapered balun, dashed-thin-line: performance of a simple coaxial coil balun without tapering but same wire length but less winding coupling.

#### ACKNOWLEDGEMENT

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