APPLICATION NOTE

Design of H.F. Wideband Power Transformers; Part II

ECO7213

Philips
Semiconductors

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CONTENTS

1 SUMMARY

In part I of this report (ECO6907) the transmission line transformer has been discussed extensively. In this second part attention has been paid to the conventional transformer as a wideband matching element in e.g. S.S.B. transmitters. The main problem in these transformers is stray-inductance. Some H.F. compensation methods are discussed and a design example is given.

2 INTRODUCTION

Report ECO6907 (Ref. 1) has been devoted entirely to the design of transmission line transformers. This type of transformer has undoubtedly the advantage of the largest possible bandwidth. However it has also some drawbacks:

1. The impedance transformation ratio is restricted to $(n : m)^2$ in which n and m are small integers
2. They are difficult to construct when a number of windings must be combined.

Therefore it is certainly worthwhile to consider the possibility of applying a conventional transformer if the abovementioned problems arise.

3 L.F. LIMITATIONS OF CONVENTIONAL TRANSFORMERS

These limitations are the same as for transmission line transformers, viz. parallel inductance and maximum flux density. They have been described in the abovementioned reports including compensation methods for the parallel inductance.

4 H.F. LIMITATION OF CONVENTIONAL TRANSFORMERS

The most important property that limits the H.F. performance of a conventional transformer is the well-known stray-inductance. The equivalent circuit of the transformer for this frequency region is shown in Fig.1.

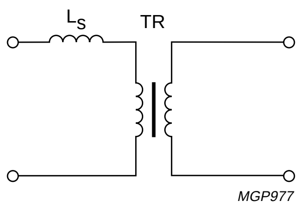


Fig.1

L_S is the stray-inductance and TR an ideal transformer. An easy way of measuring L_S is to measure the reactance between the primary terminals when the secondary is short-circuited. This can be done e.g. with the H.P. Vector Impedance meter. The most accurate result is obtained when the measurement is made at the highest frequency of operation and on the high-ohmic winding.

It is obvious that L_S must be kept as small as possible to avoid degradation of the H.F. performance of the transformer. For this end the following measures are recommended:

1. The windings must be as close to the core and to each other as possible
2. Each winding must be divided equally around the whole periphery of the core
3. Each winding must cover the core as much as possible.

Some practical steps that can be taken are:

1. The use of copper foil for the low-ohmic winding; this can be in direct contact with the core as the resistivity of the ferrite is very high. For better covering of the core two windings can be connected in parallel in such a way that one winding is wound in between the other one; the isolation material required in this case must be very thin.
2. For the high-ohmic winding enamelled copper wire can be used. An appreciable reduction of L_S can be obtained by parallel connection of two or more wires.

5 H.F. COMPENSATION OF CONVENTIONAL TRANSFORMERS

In Ref. 2 some forms of compensation have been described. They are compared in Table 1.

Table 1

NUMBER OF COMPENSATION ELEMENTS	0	1	2
Maximum X/R	0.18	0.44	1.09

Table 1 is based on a maximum input V.S.W.R. of 1.2. The quantity X/R is the reactance of the stray-inductance (referred to the primary) divided by the nominal input resistance of the transformer. Compensation with one element is possible either by connecting a capacitor in parallel with the primary winding, or by connecting a capacitor in parallel with the secondary winding.

Compensation with two elements is carried out by connecting capacitors in parallel with both primary and secondary windings. This method has already been applied several times. A short description will be given with reference to Fig.2.

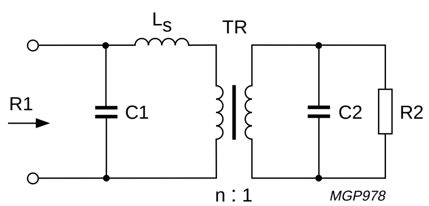


Fig.2

TR is an ideal transformer with a voltage transformation ratio of n : 1 (n > 1) and L_S is the stray-inductance (referred to the primary). R_1 is the nominal input resistance:

$$R_1 = n^2 \times R2$$

First we determine the normalized stray-inductance:

$L_{sn} = \dfrac{\omega_{max} \times L_s}{R_1}$, in which ω_{max} must be equal to or higher than 2π times the maximum frequency to be handled.

With the aid of Fig.5 we find the maximum input V.S.W.R (S) and the normalized correction capacitance (C_{1n}). Then C1 and C2 can be calculated:

$C_1 = \dfrac{C_{1n}}{\omega_{max} \times R_1}$ and $C2 = n^2 \times C_1$

A practical problem that can arise is that the capacitor across the low-ohmic winding must have such a high value that it approaches series resonance with its own lead inductance. In that case a sufficiently large number of smaller capacitors must be connected in parallel.

The practical limit to which the above described compensation system can be used is appr.:

$L_{sn} = 1$

If the stray-inductance is larger a more complicated compensation system is required. The above described system is based on a 3-element Chebyshev low-pass filter, but this can be extended to a 5-element network, which means that for the compensation 4 elements are required. This situation is depicted Fig.3.

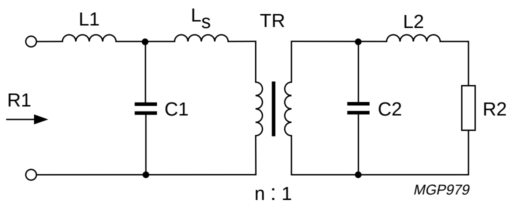


Fig.3

Two inductors (L1 and L2) have been added. The normalized elements versus the maximum input V.S.W.R. (S) have been given in Fig.6 (this graph and that of Fig.5 have been made with the aid of a computer program which is based on the design formulae for Chebyshev low-pass filters as can be found e.g. in Ref. 3.). From this graph it can be seen that L_{sn} may be as high as 1.77 for $S_{max} = 1.2$. A good practical limit is: $L_{sn} = 1.6$

The procedure for calculating the compensation elements is very similar to the previous case. We start again with determination of the normalized stray-inductance: $L_{sn} = \dfrac{\omega_{max} \times L_s}{R_1}$

With the aid of Fig.6 we find the maximum value of S and the normalized values of the correction elements C_{1n} and L_{1n}.

Now C_1 and L_1 can be calculated: $C_1 = \dfrac{C_{1n}}{\omega_{max} \times R_1}$ and $L_1 = \dfrac{L_{1n} \times R_1}{\omega_{max}}$

At the secondary side the correction components become:

$$C_2 = n^2 \times C_1 \text{ and } L_2 = \frac{L_1}{n^2}$$

In general L_2 is so small that the tracks on the p.c. board can perform this function.

6 A PRACTICAL EXAMPLE

For a low voltage S.S.B. amplifier a transformer was required with the following specification:

Frequency range: 1.6 – 28 MHz

Power handling: 52 W

Load impedance: 100 Ω

Input impedance: 4.6 Ω

The most suitable Ferroxcube material for this frequency range is 4C6. From the power handling point of view a toroid core will be chosen with the dimensions: $23 \times 14 \times 7$ mm^3, catalog nr. 4322 020 91070. The parallel reactance at 1.6 MHz measured at the secondary side must be +j400 Ω (see Ref. 1), corresponding with an inductance of 40 μH. The required number of turns is then:

$$n_{sec} = \sqrt{\frac{L \times l}{\mu_o \times \mu_r \times A}}$$

in which $\frac{A}{l} = 0.5525$ mm for this core and $\mu_r = 120$ for 4C6 material, so:

$n_{sec} = 21.9$

A logical choice would be $n_{sec} = 22$. The voltage transformation ratio is $\sqrt{\frac{100}{4.6}} = 4.66$, by which the primary number of turns must be: $\frac{22}{4.66} = 4.7$

We choose: $n_{pr} = 5$.

The required transformation ratio can be approached better if the secondary number of turns is modified from 22 to 23. Now the secondary parallel inductance becomes 44 μH and the impedance transformation ratio $\left(\frac{23}{5}\right)^2 = 21.1$ by which the input impedance will be 4.74 Ω being appr. 3% higher than the required value.

L.F. compensation of the parallel inductance can be carried out by means of a single capacitor in series with the secondary (see Ref. 1):

$$C_L = \frac{L_{psec}}{R_{sec}^2} = \frac{44 \times 10^{-6}}{100^2} = 4400 \text{ pF}$$

A standard value of 4700 pF has been chosen.

The maximum voltage across the secondary winding is: $V_{sec} = \sqrt{2R_{sec} \times P} = \sqrt{2 \times 100 \times 52} = 102$ V

Now the maximum flux density can be calculated:

$$B_{max} = \frac{V_{sec}}{\omega_{min} \times A \times n_{sec}}$$

in which A = 31.5 mm^2 for this core, so: $B_{max} = \frac{102}{2\pi \times 1.6 \times 10^6 \times 31.5 \times 10^{-6} \times 23} =$ 140 gauss at 1.6 MHz

This gives a core loss of appr. 1% or 0.5 W.

To keep the stray-inductance low the transformer has been wound as follows:

- The primary consists of the parallel connection of two windings each having 5 turns of 4 mm wide copper foil. Each winding has been equally divided around the periphery of the core; one winding was wound in between the other one. Between these two windings a thin layer of isolation material was used.
- The secondary consists of the parallel connection of two windings each having 23 turns of 0.45 mm enamelled copper wire. The method of winding was the same as for the primary.

The stray-inductance measured at the secondary side was 0.67 µH. To make the correction less critical we choose a maximum frequency of 35 MHz instead of the specified 28 MHz. The normalized stray-inductance becomes then:

$$L_{sn} = \frac{2\pi \times 35 \times 10^6 \times 0.67 \times 10^{-6}}{100} = 1.47$$

From Fig.6 we find:

S_{max} = 1.064

C_{1n} = 1.24

L_{1n} = 0.66

In this case the index 1 applies to the secondary and index 2 to the primary. The compensation elements become:

$$C_1 = \frac{1.24}{2\pi \times 35 \times 10^6 \times 100} = 56.4\ \text{pF}$$

$$L_1 = \frac{0.66 \times 100}{2\pi \times 35 \times 10^6} = 0.3\ \mu\text{H}$$

C2 = 21.1 × 56.4 = 1190 pF
L2 = 0.3/21.1 = 0.0142 µH

For C1 a standard value of 56 pF was chosen. C2 consisted of the parallel connection of 330 pF and 3 × 270 pF being 1140 pF in total; the slightly lower value was chosen because of the series inductance of the capacitors. L2 was formed by the tracks on the p.c. board.

The first measurements on the compensated transformer showed a too high V.S.W.R. at 28 MHz together with a capacitive behaviour of the impedance from which we got the impression that it was over-compensated. This appeared to be due to the capacitance between the primary and the secondary winding. Therefore we reduced the values of the compensation elements by 10 to 20%. The new values are:

C1 = 47 pF

L1 = 0.27 µH

C2 = 980 pF (parallel connection of 2 × 270 pF and 2 × 220 pF)

L2 ≈ 2 × 6 nH (tracks on p.c. board)

The complete situation is depicted in Fig.4. The load impedance Z_L consisted of the parallel connection of 2 resistors of 15 Ω, 1 resistor of 12 Ω and a capacitor of 120 pF. The latter was added to compensate the series inductance of the resistors.

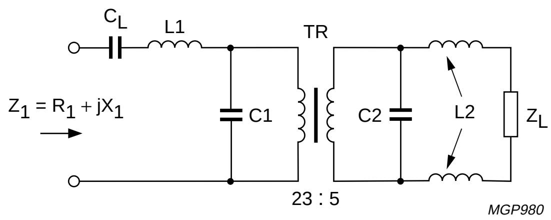


Fig.4

The results of the measurements have been summarized in Table 2.

Table 2

f (MHz)	R1 (Ω)	X1 (Ω)	V.S.W.R. (–)
1.6	98	–1.87	1.03
3.5	100.5	–2.1	1.02
7.0	98.3	–5.84	1.06
14	89.9	–4.41	1.12
20	87.4	+1.24	1.14
28	100	+7.18	1.07

It can be seen from Table 2 that the maximum input V.S.W.R has been reduced to 1.14. Without H.F. compensation this V.S.W.R. would have been appr. 3 at 28 MHz.

7 REFERENCES

1. A.H. Hilbers; *“On the Design of H.F. Wideband Power Transformers”*, C.A.B. report nr. ECO6907, March 14, 1969.
2. H. Nielinger; *“Optimale Dimensionierung von Breitbandanpassungsnetzwerken”*, N.T.Z. 1968, Heft 2, pp. 88 to 91.
3. S.B. Cohn; *“Direct-Coupled-Resonator Filters”*, Proc. IRE, February 1957, pp 187 to 196.

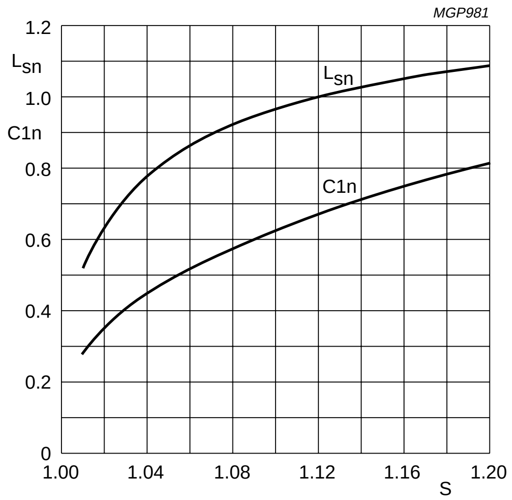
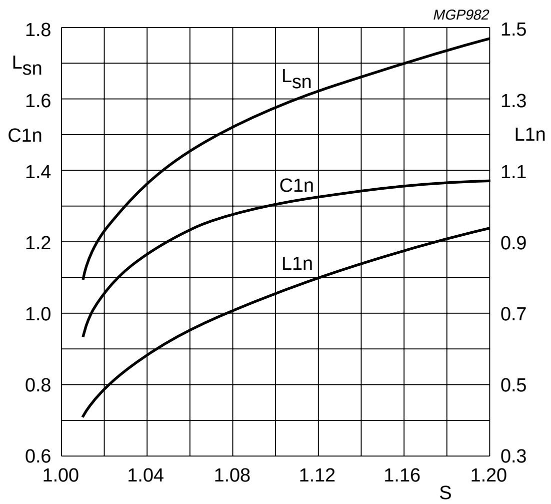

MGP981
1.2
L_{sn}
1.0
C1n
0.8
0.6
0.4
0.2
0
1.00
1.04
1.08
1.12
1.16
1.20
S
L_{sn}
C1n

Fig.5


MGP982
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L_{sn}
1.6
C1n
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1.2
1.0
0.8
0.6
1.5
1.3
L1n
1.1
0.9
0.7
0.5
0.3
1.00
1.04
1.08
1.12
1.16
1.20
S
L_{sn}
C1n
L1n

Fig.6