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Internal Broadband Antennas for Digital Television Receiver in Mobile Terminals

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***Abstract*—The Implementation of internal broadband antennas for digital television receiver (DTV) in mobile terminals operating at the lower UHF band is studied comprehensively. Certain challenges such as inherently narrow impedance bandwidth of electrically small antennas and limited volume available for the antennas inside mobile terminals are identified and handled. The proposed design principle for DTV antennas is to decrease the total efficiency of the antenna to a level which is just good enough to ensure a sufficient signal-to-noise ratio and that way make the size of the antenna sufficiently small. The limits for the size of broadband capacitive coupling element-based DTV antenna structures inside handsets of different sizes are studied. In the end, a manufactured prototype and simulated design are presented and compared with the limits studied earlier in the paper. The results show that the studied antenna concept is a promising candidate for broadband DTV antennas in mobile terminals. The work also increases general understanding on the implementation of antennas based on the radiation of the finite ground plane.**

***Index Terms*—broadband antennas, broadband communication, broadcasting, digital TV, DVB-H, impedance matching, microstrip antennas, mobile antennas, receiving antennas.**

I. Introduction

RECENTLY, there has been a significant increase in the number of different functions and radio systems built into handheld devices. In addition to the traditional cellular systems, many other radios have been introduced in mobile terminals, such as FM radio, digital television (DTV), 3 G, GPS, Bluetooth, and WLAN. The size of the terminal has remained roughly the same despite the increased number of radio systems and at the same time internal antennas are preferred. Thus, the volume available for antennas is very limited inside a mobile terminal and the size of the antennas is a very critical issue.

The wavelength of DTV at the UHF frequencies (0.47–0.75 GHz in mobile terminals which include also E-GSM) is 400–640 mm, and as the typical handset size is 100–130 mm, an internal DTV antenna inside such a terminal is electrically rather small. Due to the physical limitations [1]–[4], the implementation of small broadband DTV antennas is very challenging with the typical matching criterion (6 dB return loss) of current cellular antennas.

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The available internal broadband DTV antenna solutions can be divided into two main categories according to the instantaneous frequency bandwidth. The first group is antennas with fixed broadband matching covering the whole band. The main challenge is to cover the whole band with sufficient performance. Broadband solutions are introduced e.g., in [5]–[7]. The other main group is formed by electrically tunable antennas which create an instantaneous resonance at a suitable frequency so that at least a single 8-MHz channel is covered. Electrically tunable antennas are introduced e.g., in [8], [9] and [10]. One challenge is the non-linearity caused by the semiconductor tuning component, which becomes a problem during the simultaneous use of E-GSM transmission when a part of the transmitted E-GSM signal is coupled to the DTV antenna. In addition, the tuning circuit needs a control voltage, might consume valuable power, and increases the complexity of the whole DTV receiving system. According to [11] another challenge arises when the terminal is doing a handover from one transmitter to another. While receiving signal from the first transmitter, the receiver needs to simultaneously make a new channel-scan, which might be complicated to perform with a single tunable antenna.

Due to the inherently narrow impedance bandwidth of small antennas, tunable antennas seem like a reasonable choice for DTV antennas, but on the other hand the high linearity and fixed matching of passive implementation with a simple antenna structure motivates the use of broadband DTV antennas. It is also expected that the effect of the user on the matching is smaller on broadband antennas. Therefore, this paper concentrates only on implementation and characteristics of such antennas.

The ground planes of the printed circuit board (PCB), EMC shieldings and other conductive structures (such as the display) of a handheld device create a solid RF counterpoise, here called a chassis. This electrically conductive chassis has a significant effect on the antenna operation because it operates as the main radiator at lower UHF frequencies, i.e., below 1 GHz [12]. Therefore, the size of the chassis has a significant effect on the achievable bandwidth and also on the size of the antenna element [12], [13], especially at the DTV frequencies [14]. In this work the chassis is expected to be a solid piece of metal and thus any direct feed-based antenna structures as presented e.g., in [6], [7] are not considered.

To the authors' knowledge there exist no scientific publications with a systematic study of implementation of internal broadband DTV antennas. The purpose of this paper is to present a thorough study of the implementation of such DTV antennas in mobile terminals. First, the used antenna concept, capacitive coupling element structures, are briefly introduced.

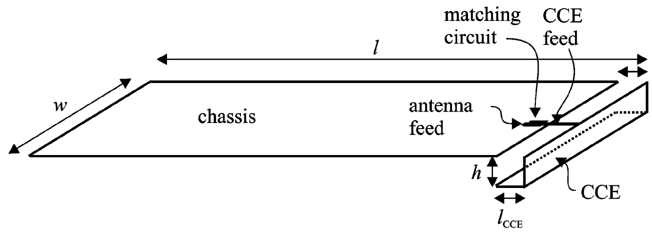

Fig. 1. Capacitive coupling element (CCE) structure.

After that the essential design and performance parameters are derived and discussed. In this paper DVB-H (digital video broadcasting—handheld) is used as a DTV standard since it is a typical broadband DTV standard for mobile terminals and most probably used at least in Europe. Next, the design of a simplified full-metal DTV antenna is presented. Minimum height and volume required of a DTV antenna to reach a certain performance inside mobile terminals of different sizes are presented. Some methods to reduce the thickness of the antenna are also handled. In the end the results are compared and verified with a prototype and simulated designs.

II. Capacitive Coupling Element Antenna Structures and Matching Circuits

Traditional mobile terminal antennas, such as PIFA, create the antenna resonance and couple currents to the surface of the chassis [12], [13]. In principle, traditional terminal antennas could be used as DTV antennas but the bandwidth would be rather narrow or the antenna volume would be too large to be placed inside a handset. Since the antenna element itself is only a minor radiator below 1 GHz, the volume occupied by the element can be decreased significantly by introducing capacitive coupling elements (CCE) whose principal function is only to couple currents to the chassis [12], [13]. The resonance is then created with a separate matching circuitry outside the coupling element e.g., using well-known basic matching methods presented in text books, see e.g., [15], [16]. One possible capacitive coupling element antenna structure is shown in Fig. 1.

In order to achieve the largest possible impedance bandwidth, the location and shape of the capacitive coupling element need to be chosen in such a way that the coupling between the coupling element and the chassis dominant wavemode is maximized [13]. For the strongest possible coupling, the coupling element needs to be placed beyond the edge of the chassis with the element bent over the shorter edge of the chassis, see Fig. 1. The vertical part of the element could be further lengthened on the upper side of the chassis but that place is typically reserved for other components (such as connectors, buttons, microphone/earpiece, camera, etc.) in real terminals.

For the structure shown in Fig. 1 it is shown in [17] that with constant distance d the maximum impedance bandwidth at the DTV frequencies is achieved when l_{CCE} equals d. Further increase in the bandwidth can only be achieved at the expense of larger volume occupied by the coupling element, i.e., increasing the length of the element l_{CCE} and/or the height h of the CCE from the upper surface of the chassis. Since DTV is a receive-only-system the specific absorbtion rate (SAR) values can be neglected.

The matching circuit can be implemented using lumped or distributed technologies [5],[13], [17]. Below 1 GHz the advantage of high-Q lumped elements (such as inductors and capacitors) over distributed elements is the significantly smaller required matching circuit area [17]. According to [13], the matching circuit might be integrated into the RF front-end module in commercial mobile terminals. Another option is introduced in [18] where the coupling element and the matching circuit are integrated into a single component using the LTCC technology.

The bandwidth can be increased with a multi-resonant matching circuit, which can be implemented with coupled high-Q resonators, see e.g., [19]. In practice the number of matching resonators in a matching circuit is limited by design complexity, losses and the fact that the bandwidth increase from each additional resonator gets smaller [20]–[22]. However, the matching circuit needs to be optimized for each capacitive coupling element and chassis combination [13].

III. Design Parameters of Broadband Internal DTV Antennas

A. Performance of DTV Receiving Antennas

A basic requirement for radio systems is to reach a sufficient signal-to-noise ratio over the frequency band of operation. In order to minimize the size of the antenna element, the design principle of DTV antennas is to provide a performance which is just enough for guaranteeing the operation with a certain reliability level.

As was discussed, it is difficult to cover the whole DTV frequency band with a good matching level (e.g., 6 dB return loss). As is commonly known, the impedance bandwidth of an antenna can be increased by sacrificing a part of the total efficiency [4]. Even though the method is not commonly recommended, in small internal DTV antennas the total efficiency needs to be sacrificed in order to make the size of the antenna feasible for today's mobile terminals [23], [24]. This is possible since DTV is a receive-only-system and lower total efficiency compared to transceiver antennas can be accepted [25]. In order to estimate the lowest required total efficiency of a DTV antenna, an equation is derived in [26]:

$$\eta_{tot,min} = \frac{4\pi P_{rec,min}\eta_0}{E_{inc}^2 c_0^2 D} f^2 \tag{1}$$

where $P_{rec,min}$ is the sensitivity of the receiver, η_0 is the far field wave impedance in the free space 377Ω, E_{inc} is the typical minimum electric field strength guaranteed by the broadcasting network, c_0 is the speed of light, D is the directivity of the antenna and f is the frequency used. According to [23], the sensitivity of a typical DTV (DVB-H) receiver can be expected to be at least $P_{rec,min} = -90$ dBm. In [27] it has been reported that the minimum electric field strength in the DTV indoor reception is typically better than $E_{inc} = 55\ \mathrm{dB}\mu\mathrm{V/m}$. The directivity is about 2 dBi for small antennas [28]. These roughly approximated values and (1) yield that the minimum total efficiency required in a DTV antenna is in the order of $-16\cdots-12$ dB over the band.

The expected DTV antenna performance in the DVB-H standard followed in this paper has been given in terms of realized gain which consists of the directivity and the total efficiency [29]. In the DVB-H system specifications the realized gain of an antenna inside a mobile terminal is expected to be in the order of -10 dBi at 0.47 GHz and it increases linearly in dB's to about -6.5 dBi at 0.75 GHz. When the directivity (2 dBi) is excluded, the total efficiency is expected to be in the order of $-12\cdots-8.5$ dB according to the above-mentioned realized gain. Concluding the calculations above, the expected performance is at least 3.5 dB higher than the estimated minimum required total efficiency in the previous paragraph. In the following the presented realized gain limit is considered as the performance specification for internal DTV antennas. The specification is given for an antenna inside a real terminal [23], [24]. In this work the antenna structures include only metal chassis (and in the prototype low-loss printed circuit board), antenna element and matching circuit. Thus, an extra margin to the specification is needed in the designs in this work because in the final commercial products the printed circuit board (typically FR4), plastic covers, display, battery, earpiece, microphone and other lossy parts of the terminal cause additional losses.

B. Broadening Impedance Bandwidth of DTV Antennas

An inherently narrow-band antenna can be matched in such a way that the impedance bandwidth is significantly increased at the expense of the total efficiency [21], [29], see Fig. 2. One can understand the idea considering that the available area below the narrow-band matching curve is redistributed over a larger impedance bandwidth. However, one should note that the areas below the narrow-band and broadband matching curves in Fig. 2 are not exactly equal because the radiation quality factor of the antenna changes as a function of the frequency. The remaining questions are how much and which way the total efficiency can be sacrificed in a DTV antenna. There are basically three options. Firstly, one can use resistive loading of the antenna (lowering radiation efficiency), secondly, one can use resistive matching (attenuators) or, thirdly, one can accept higher mismatch between the antenna and the receiver (lowering matching efficiency). Comparing these three cases, the third option can be shown to provide the widest bandwidth for a given decrease of the total efficiency [17].

The radiation efficiency is close to 100% for a low-loss full-metal (only chassis and antenna element) antenna structures. In that case the only parameter affecting the realized gain is the return loss of the antenna. Now, the realized gain can be plotted at the edge of the impedance band as a function of the return loss matching criterion, see Fig. 3. As can be seen, in the low-loss case the matching of 1.5 dB return loss yields a realized gain of approximately -3.5 dBi. Since the realized gain specification is between -10 and -6.5 dBi over the DTV band, the minimum margin to the specification is at least 3 dB, which can be reserved for the roughly approximated implementation losses introduced by the other components in practical antennas. Thus, 1.5 dB return loss is used in this work as the design parameter for the ultimate lowest acceptable matching criterion for low-loss internal DTV antenna designs. As is well known, the above-mentioned implementation losses consequently improve the matching level and thus the matching of antenna structures of complex commercial mobile terminal and metallic low-loss antenna structures studied in this paper are generally not directly comparable. One should also note that the used 3-dB margin to the realized gain specification in the low-loss case is just one possible choice, and based on the typical total efficiencies reported for cellular antennas of commercial terminals [26], [30], [32].

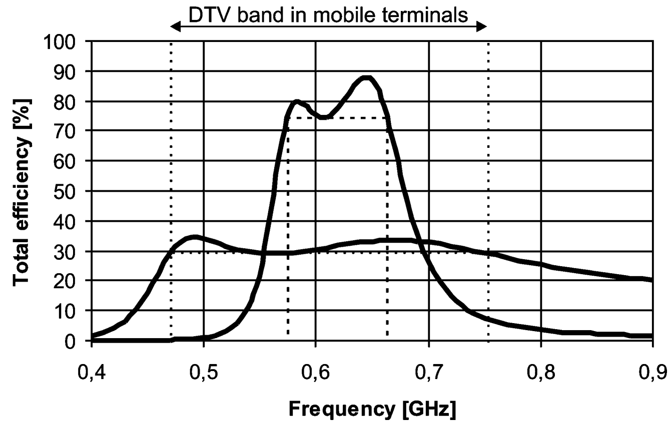

Fig. 2. Conceptual graph on how a narrow-band antenna can be matched so that a larger bandwidth is covered at the expense of the total efficiency.

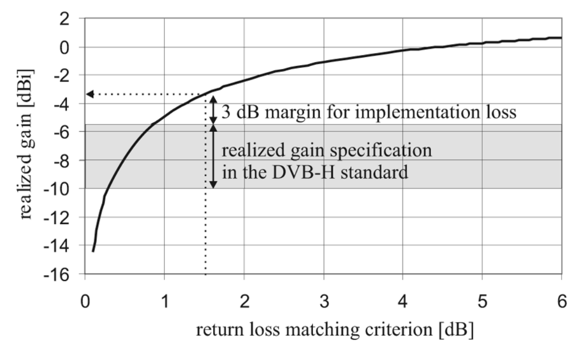

Fig. 3. Realized gain estimation of a DTV antenna as a function of the return loss matching criterion.

IV. Design of Broadband DTV Antennas With Capacitive Coupling Elements

A. Simulated Example Design

The purpose of this section is to demonstrate a simplified low-loss capacitive coupling element-based DTV antenna design for a typical-size monoblock terminal. An example antenna is designed with dimensions $l = 110$ mm, $w = 48$ mm and $d = l_{CCE} = 5$ mm, see Fig. 1. The height h of the antenna element from the surface of the chassis is a variable. The input impedance of the antenna structure (without the matching circuit) with several values of height h (0.25 mm steps) against 50 Ω reference impedance was simulated from the CCE feed point across 0.3–1.5 GHz with a method of moments-based electromagnetic simulator IE3D [32]. The impedance curves for certain cases ($h = 5$, 10 and 15 mm) are shown on the capacitive half of the Smith chart in Fig. 4. As can be seen, the input resistances R are fairly low ($R < 15\ \Omega$) and the reactances X are highly capacitive ($X = -75\cdots-200\ \Omega$) at the DTV

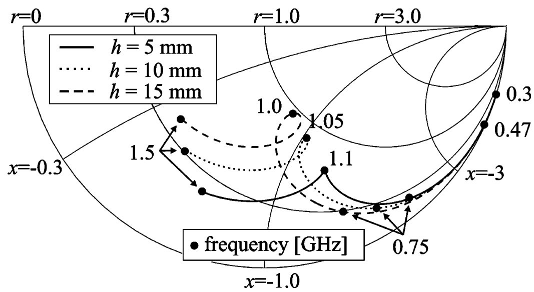

Fig. 4. Impedance curves at 0.3–1.5 GHz when h is 5, 10, and 15 mm.

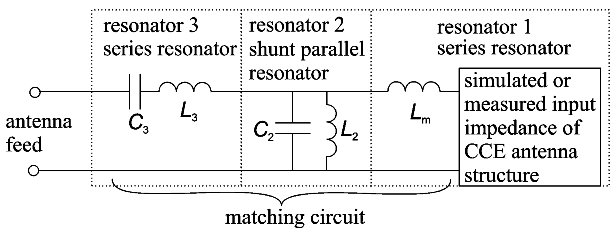

Fig. 5. General triple-resonant matching circuit with series and parallel resonators in turn.

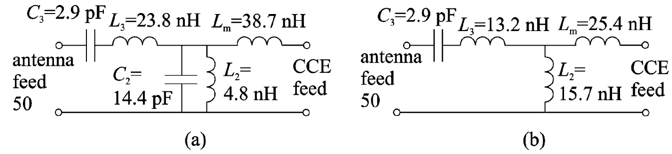

Fig. 6. (a) Triple-resonant and (b) dual-resonant matching circuits and their component values for the example design.

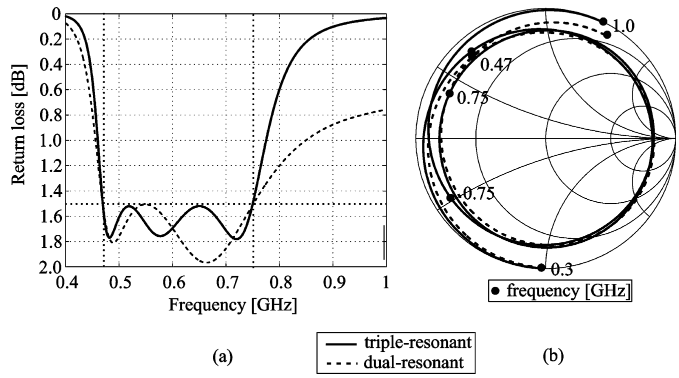

Fig. 7. Simulated (a) return loss in the Cartesian coordinate system and (b) the complex impedance on the Smith chart for the example design.

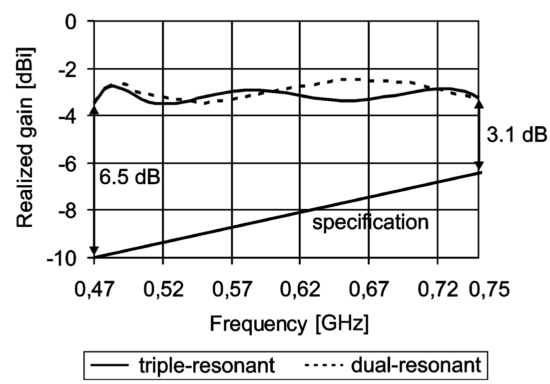

Fig. 8. Simulated realized gains of the example design.

frequencies. One can also note the lowest order wavemode/resonance of the chassis at 1.1, 1.05 and 1.0 GHz for $h = 5$, 10 and 15 mm, respectively [12]. Unfortunately the resonances are out-band for the DTV system frequencies and they can not be exploited in the matching of the antenna.

The next task is to match the antenna across the DTV band with a multi-resonant matching circuit. In this work dual-resonant and triple-resonant matching circuits are used. With more additional resonators the bandwidth increase gained from each resonator saturates and also in practice the matching circuit becomes rather complex. In [33] it is reported that a capacitive coupling element antenna structure is inherently a series-type resonator and thus, the first resonator in the combination of the antenna and the matching circuit is series-type. Since CCE structures are highly capacitive at the DTV frequencies (see Fig. 4), the resonant frequency of the first resonator can be tuned to the DTV frequencies with a rather large-value series inductor. In order to achieve the lowest possible complexity of triple-resonant matching circuit (five matching circuit components), the second resonator is a shunt parallel resonator and the third is again a series resonator, see Fig. 5 [19].

Dual-resonant matching circuits can be implemented either by leaving out the third resonator inFig. 5 (three matching circuit components) or by using coupled series-type resonators, in which the impedance inverters can be partly integrated into resonators, as presented in [35] (four matching circuit components).

The inductance value L_m of the first inductor in the triple-resonant matching circuit (see Fig. 5) can be estimated rather easily. The input reactance at the center frequency f_c of the DTV band is $X \approx -150\ \Omega$ (see Fig. 4) and thus the inductor value L_m can be estimated from equation $\omega L_m \approx 150\,\Omega$, where $\omega = 2\pi \cdot f_c$ and thus $L_m \approx 39$ nH. Even though the theory for general multi-resonant matching circuits is known [20]–[22], there exist no closed-form design formulas and thus the rest of the matching circuit needs to be designed with an automatic optimization tool of a circuit simulator. In this work APLAC [22] was used. The matching circuits were tuned to be optimal from the bandwidth point of view [22]. At this stage ideal (i.e., lossless) lumped elements were used in the matching circuits.

The height h of the coupling element was chosen to be the smallest possible which enables at least 1.5 dB return loss with a low-loss full-metal antenna structures across the DTV band. (The effect of the distance d is to be handled later.) However, the smallest possible heights are surprisingly high, $h = 14.5$ mm and $h = 18.5$ mm with triple-resonant matching and dual-resonant matching, respectively. Thus the volumes of the coupling elements are $3.5\ \mathrm{cm}^3$ and $4.4\ \mathrm{cm}^3$. The corresponding ideal matching circuits are illustrated in Fig. 6. The return losses and realized gains against 50 Ω reference impedance for both studied cases are shown in Figs. 7 and 8.

On the Smith chart in Fig. 7(b) the reflection coefficients have in both cases the inner loops circulating the center of the Smith chart symmetrically indicating optimal frequency response from the bandwidth point of view [22]. The realized

TABLE I
TYPICAL TOLERANCES OF THE MURATA CHIP INDUCTORS AND CAPACITORS

tolerances of Murata chips	small component value (abs. nH/pF)	large component value (per cent)
best inductors LQP15M series	0.1	2
worst inductors LQW04A series	0.5	5
typical capacitors GJM03 series	0.25	5

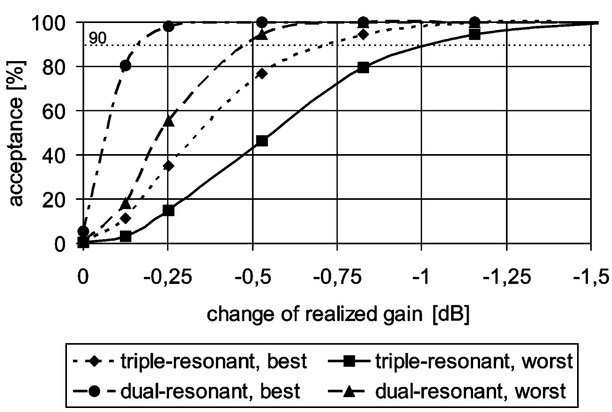

Fig. 9. Effect of the variation of the matching circuit component values on the realized gain.

gains are on average -3.5 dB which agrees very well with the estimation presented in Fig. 3. The minimum margin to the realized gain specification in both cases is 3.1 dB, which can also be expected for this low-loss design according to Fig. 3. At lower DTV frequencies the margin is much larger than 3.1 dB, even 6.5 dB, which is actually very useful because at the lower DTV frequencies the radiation resistance of the CCE antenna structure is rather low (see Fig. 4) and thus the losses in real antenna structures will decrease the radiation efficiency more than at the upper DTV frequencies where the radiation resistance is larger. The simulated antenna structure and matching circuit provide a good starting point for the implementation process of a more realistic DTV antenna.

Next, the change of the realized gain due to the typical statistical variation of the matching circuit component values is studied. The analysis is performed for both studied cases. The typical tolerances of inductance and capacitance values have been obtained for Murata high frequency chip components in [35] and they are shown in Table I. "Small" component value means smaller than 10 nH for inductors and smaller than 5 pF for capacitors. Because of lots of differences in the tolerances of the available Murata inductor series, the "best" inductors mean the tightest tolerance available and the "worst" inductors vice versa. For capacitors the tolerances are typically the same independent of series.

The analysis is performed in such a way that the effect of the statistical variation of the matching circuit component values on the return loss is calculated with the Monte Carlo analysis tool in APLAC. After that the change of the realized gain is calculated from the change of the return loss. The results are shown in Fig. 9. The results can be interpreted the following way: If we tolerate e.g., -1 dB maximum change of the realized gain, 90% of the products (acceptance level 90%) in the worst triple-resonant case will have less than -1 dB change in the realized gain. At the same time if we use dual-resonant matching with the 90% acceptance level, we have the maximum change of the realized gain of only -0.45 dB. On the other hand, we have also 4 mm thicker antenna element than in the triple-resonant case.

As can be seen, first of all, the dual-resonant and triple-resonant cases have surprisingly large difference in the acceptance levels with a given change of the realized gain although in the triple-resonant matching circuit there is only one shunt capacitor more (C_2) than in the dual-resonant matching circuit. Secondly, there is a big difference in the acceptance levels between the best and worst tolerance inductors. In the end, the decision of the used components and topology of the matching circuit are to be dictated by the size and performance of the antenna, and also the cost of the components.

TABLE II
TERMINAL DESIGNATIONS AND DIMENSIONS.

terminal name	dimensions l x w [mm x mm]
small monoblock	100 x 44
typical monoblock	110 x 48
large monoblock	120 x 52
tablet	135 x 75

B. Height and Volume of the Capacitive Coupling Element Versus Handset Dimensions

This section provides information about the height and volume occupied by the coupling element in different size terminals. As discussed, the size of the terminal, which determines the size of the chassis, affects significantly the size of the capacitive coupling element. When reducing the terminal dimensions we need a larger antenna element to obtain the same performance and vice versa [12]–[14]. Hence, there is a motivation to study what is the smallest possible antenna element with given outer dimensions of a terminal. In this paper four different terminal dimensions (l and w in Fig. 1) are used, three of them are for a monoblock terminal and one is for a tablet-size terminal, see Table II. Other terminals types are not considered here because the bar-type (monoblock or tablet) terminal is considered to be the most challenging case and the mechanics is also pretty well known. For example in an open clamshell the total length of the chassis is rather long (even 160 mm) and thus, the implementation of a DTV antenna within a small volume is less problematic [14].

The minimum height h of the capacitive coupling element enabling at least 1.5 dB return loss is studied for each of the four terminal sizes the same way it was done in the previous subsection. The minimum outer dimensions (l, w and h) of the terminal are supposed to enclose the chassis, antenna element and other relevant components of the antenna structure, see Fig. 10.

The matching circuit topologies are similar to those presented in Fig. 6, only the component values are slightly different in each case. In principle, the S-parameters of real matching circuit components could be used in the simulations but non-idealities

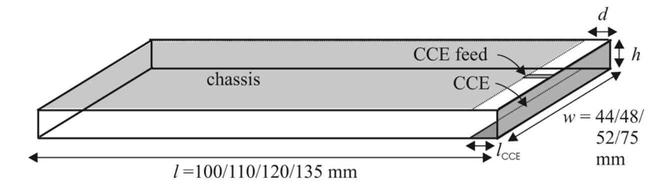

Fig. 10. Minimum outer dimensions of the different-size terminals. Gray color illustrates the capacitive coupling element and the chassis. The sketch is not in the real scale.

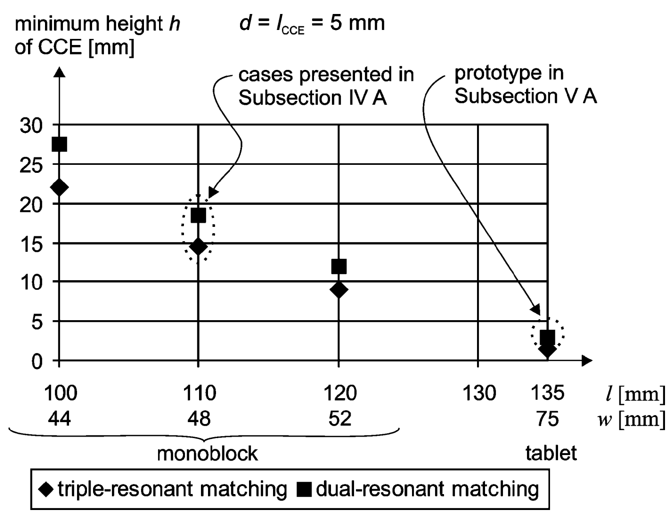

Fig. 11. Effect of different terminal sizes on the minimum height of the coupling element in dual-resonant and triple-resonant matching cases. $d = l_{CCE} = 5\ P\mathrm{mm}$.

TABLE III
VOLUME [CM^3] OCCUPIED BY THE ANTENNA ELEMENTS IN THE DUAL-RESONANT AND TRIPLE-RESONANT MATCHING CASES AND THE AVAILABLE GROUNDED AREAS [CM^2] OF THE CHASSIS $d = l_{CCE} = 5$ MM

terminal name	dual-resonant	triple-resonant	chassis area
small monoblock	6.1	4.7	41.8
typical monoblock	4.4	3.5	50.4
large monoblock	3.1	2.3	59.8
tablet	1.1	0.56	97.5

such as losses, discrete component values and parasitics would make the comparison of the results very difficult and actually inconsistent. As a result of the simulations the minimum height h of the antenna element is plotted as a function of the length l of the terminal in Fig. 11 and the volumes occupied by the element and the available grounded areas of the chassis in different cases are reported in Table III.

Based on the results in Fig. 11 and Table III, as expected, the size of the chassis has generally very significant effect on the size of the antenna element. While very big antenna elements are needed to reach the 3-dB margin to the realized gain specification with fairly short monoblock terminals, the tablet-size terminal can allow a rather small antenna element. This is because the radiation quality factor of the chassis increases when its size (both length and width) is decreased and thus the available bandwidth of the combination of the (unchanged) antenna element and the chassis is reduced. In order to maintain the available bandwidth, a smaller chassis can be compensated with a

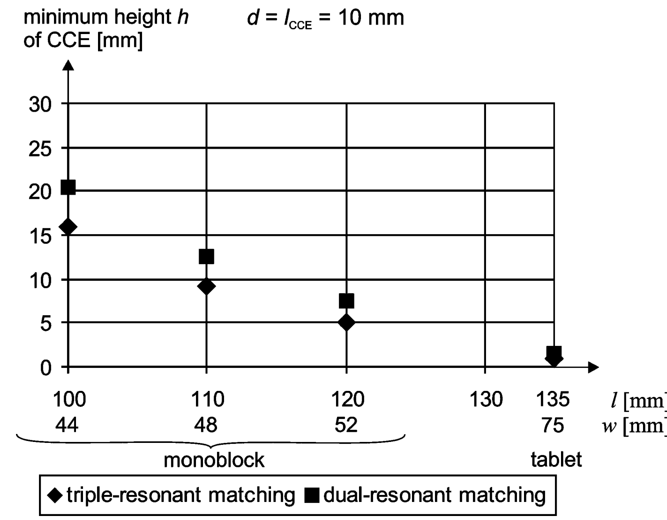

Fig. 12. Effect of different terminal sizes on the minimum height of the coupling element in dual-resonant and triple-resonant matching cases. $d = l_{CCE} = 10\ \mathrm{mm}$.

TABLE IV
VOLUME [cm^3] OCCUPIED BY THE ANTENNA ELEMENTS IN THE DUAL-RESONANT AND TRIPLE-RESONANT MATCHING CASES AND THE AVAILABLE GROUNDED AREA [cm^2] OF THE CHASSIS $d = l_{CCE} = 10\ \mathrm{mm}$

terminal name	dual-resonant	triple-resonant	chassis area
small monoblock	9.0	7.0	39.6
typical monoblock	6.0	4.4	48.0
large monoblock	3.9	2.6	57.2
tablet	1.1	0.75	93.8

larger antenna element which has both smaller radiation quality factor and increased coupling to the chassis lowest order wavemode [12]. Thus, the more the chassis size is decreased, the more the antenna element size needs to be increased. Triple-resonant matching instead of dual-resonant matching is shown to decrease the minimum height of the coupling element several millimeters, especially in smaller terminals. This can also be expected since in the triple-resonant case a smaller antenna element is enough to guarantee sufficient bandwidth compared to the dual-resonant case.

Nevertheless, the heights h of e.g., 15 mm of the antenna elements derived in the previous paragraph are too big for today's commercial monoblock terminals. As discussed in Section II, the bandwidth of the coupling element antenna can be increased by increasing the distance d of the coupling element. Vice versa, if the bandwidth of the antenna is kept equal, the increased d makes it possible to decrease the height h of the antenna element. In the following the minimum height h of the coupling element has been studied with the antenna structure having $d = l_{CCE} = 10\ \mathrm{mm}$ (see Fig. 1). One should now notice that since the total length of the whole structure is held constant, the chassis becomes 5 mm shorter than in the $d = l_{CCE} = 5\ \mathrm{mm}$ case. The procedure has been similar to the previous study. The results are illustrated in Fig. 12 and Table IV. As can be seen, the same trend can be noticed as for the $d = l_{CCE} = 5\ \mathrm{mm}$ case. When compared to the previous case, the height of the coupling element can now be made much smaller at the expense of larger total volume occupied by the element and the slightly smaller available grounded area of the chassis.

C. Capacitive Coupling Element Height Versus Performance of the Antenna

The antenna elements even with a height of only 9 mm (see Figs. 11 and 12) are still fairly big for today's terminals. As presented, while the performance is maintained, the height of the antenna element can be made smaller by making the distance d of the coupling element larger. However, in current mobile terminals the distance d and the volume reserved for antennas is very limited and the volume of the antenna cannot be increased very much. Thus, obviously the only way to make the height of the antenna element smaller with the used antenna concept is to further sacrifice a part of the performance of the antenna. Earlier it was presented that worsening the return loss is the most efficient way to increase the bandwidth (or consequently decrease the size of the antenna element) with a given decrease of the total efficiency. However, the choice made in this publication was that 1.5 dB return loss is the ultimate lowest acceptable matching level and thus the return loss will not be worsened anymore. Therefore, if we want to make the height of the antenna element smaller and compensate the inherently narrowed bandwidth, we are required to have additional resistive losses in the antenna structure, i.e., lower radiation efficiency. That can be done either by using lossier materials/components or by placing a small resistor between the antenna structure and the matching circuitry. In this work we use the latter method because it is a more systematic way to control the additional losses. One should note that the resistance of the resistor used here is a design variable and it can be partly or totally omitted in real antennas because the relatively large series inductor (L_m, see Fig. 6) in the matching circuit contains a series resistance in its equivalent circuit [35].

Two cases are to be handled in this study: typical-size monoblock in which the distance d is 5 or 10 mm with the triple-resonant matching. The starting values for the coupling element heights are $h = 14.5$ mm for the $d = 5$ mm case and $h = 9.25$ mm for the $d = 10$ mm case, see Figs. 11 and 12. The height h of the element is decreased using 2 mm steps. The smallest possible resistor, which yields 1.5 dB return loss matching over the DTV band is used. The realized gain is also simulated for each case and the margin to the specification is calculated. As a result of the study the minimum margin to the realized gain specification is plotted as a function of the height h of the coupling element, see Fig. 13.

As can be seen, the height h of the antenna element can be decreased several millimeters by sacrificing the minimum margin 3 dB by less than 1 dB. For example in the $d = 10$ mm case we can drop the height h from 9.25 mm to 6 mm by sacrificing only 0.3 dB or to only 4 mm by sacrificing 0.6 dB. After that the minimum margin drops quite quickly, though.

V. Antenna Prototypes and Simulated Designs

This section presents a certain capacitive coupling element-based antenna prototype and simulated design. The results will be compared with those of Section IV.

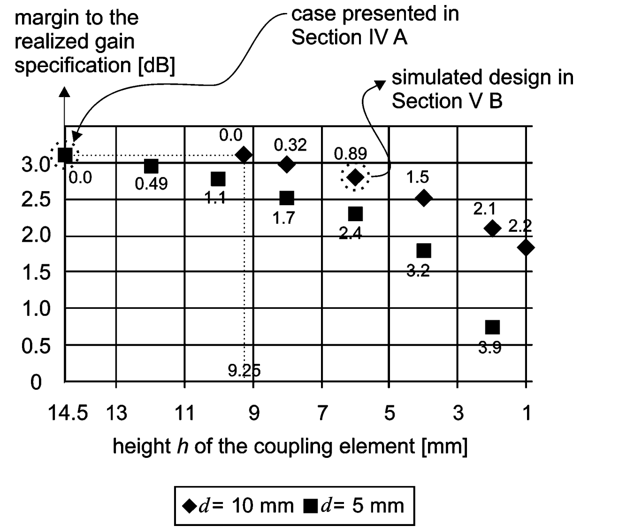

Fig. 13. Minimum margin to the realized gain specification versus the height of the element (triple-resonant matching) for typical-size monoblock. x.x Ω is the resistance of the resistor added between the antenna element and the matching circuit.

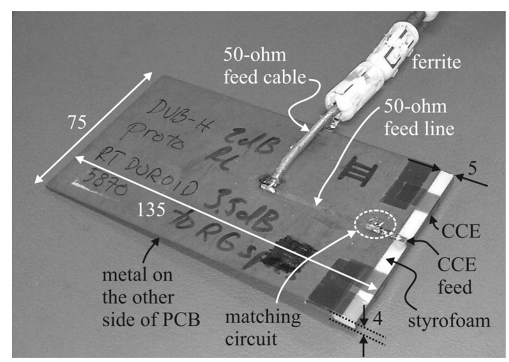

Fig. 14. Photograph of the tablet-size DTV prototype. The shape and dimensions (in mm) of the antenna are similar to shown in Fig. 10.

A. Manufactured Prototype for a Tablet-Size Terminal

This section introduces a manufactured and measured DTV antenna prototype intended for a tablet-size mobile terminal. The dimensions of the coupling element are $d = l_{CCE} = 5\,\mathrm{mm}$, $w = 75\,\mathrm{mm}$ and $h = 4\,\mathrm{mm}$, see Figs. 10 and 14, and thus the total volume occupied by the element is only $1.5\ \mathrm{cm}^3$. The prototype is implemented on low-loss Rogers RT Duroid 5870 printed circuit board. The matching is implemented with a dual-resonant lumped-element matching circuit, see Fig. 15. The inductors in the matching circuit are from Murata's LQW18A series, the capacitor is from Murata's GRM18 series [35] and the components are modeled in simulations with the S parameter available in [36]. The simulated and measured return loss are shown in Fig. 15. They are at least 2 dB across the DTV band. The difference between the simulated and measured results is expected to be caused by the slight differences in the simulation model and the manufactured prototype and the statistical variation of the matching circuit component values from the nominal values. As discussed in Sections III-A and III-B, in real terminals the implementation losses would further improve the return loss at the expense of the realized gain. The realized gain of the

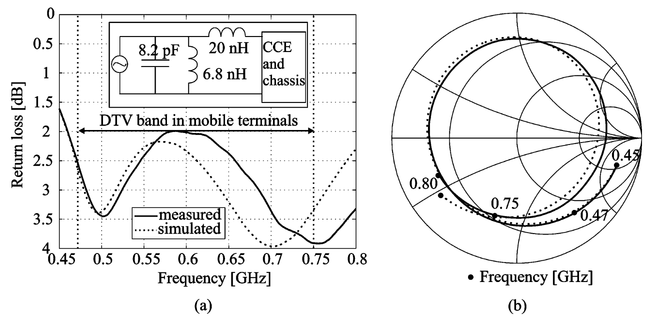

Fig. 15. Simulated and measured return loss of the tablet-size prototype (a) in the Cartesian coordinate system and (b) on the Smith chart.

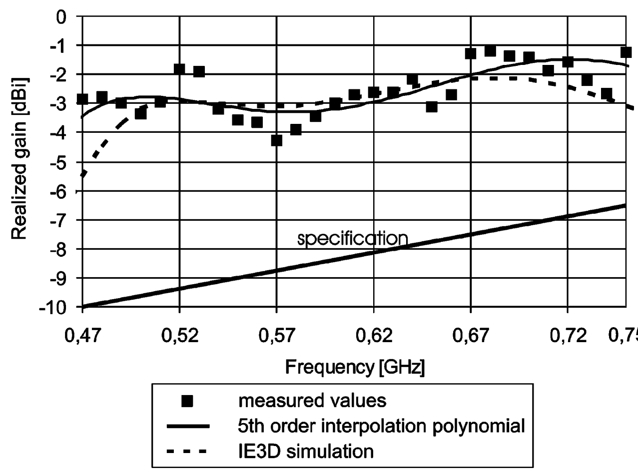

Fig. 16. Simulated and measured realized gain of the tablet-size DTV prototype.

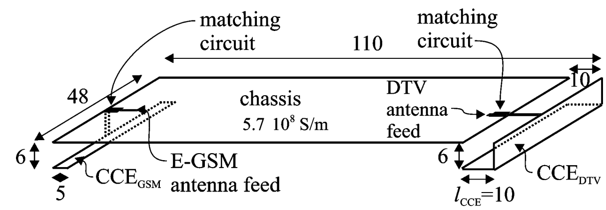

Fig. 17. Structure of the simulated design for typical-size terminal. Dimensions are in mm.

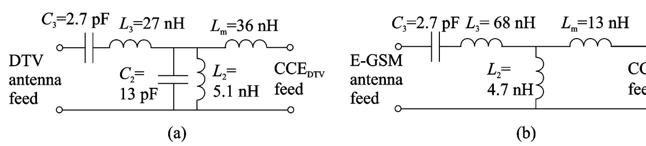

Fig. 18. Matching circuits for simulated (a) DTV and (b) E-GSM antennas.

prototype was measured with the Satimo Stargate antenna measurement system [37] for which the measurement inaccuracy is ± 1 dB given by the manufacturer. Due to the ripple the measured results are mathematically interpolated with a 5th order polynomial. The simulated and measured realized gain is shown in Fig. 16. Across the DTV band the minimum margin to the realized gain specification is 3.5 dB. Because the chassis is the main radiator, the far field directional pattern is similar to that of a dipole antenna and the directivity in the direction of the main lobe is 2 dBi across the DTV band. Thus, the total efficiency is 2 dB lower than the realized gain in Fig. 16, and in the minimum the (simulated) total efficiency is -7.5 and -5 dB at 0.47 and 0.75 GHz, respectively. However, the total efficiency is at least 7 dB larger than the minimum required total efficiency estimated in Section III-A. The far field is practically linearly polarized, i.e., the long axis of the chassis oriented parallel to the z-axis generates the far field electric field parallel to the elevation (theta) unit vector of the standard spherical coordinate system. The minimum polarization discrimination in the main lobe is 22 dB, and not very large improvement can be expected at the frequencies lower than 1 GHz with handheld-size terminals [23]. Thus, in the worst case rather low polarization efficiency can be expected.

According to Fig. 11 the minimum height h of the coupling element is 3 mm in the lossless dual-resonant case, which covers 0.47–0.75 GHz. In the manufactured prototype the height of the coupling element is selected to be 1 mm larger, i.e., 4 mm, due to practical reasons. Firstly, due to the effect of the user, the usable bandwidth is typically downtuned [14], [38]. Thus, it is useful if there is some extra bandwidth available above 0.75 GHz. Secondly, in real antennas the matching circuit cannot be designed as easily as with ideal components. Thus, some design margin is required to compensate for the effect of non-idealities such as discrete component values and the statistical variation of the component values from the nominal ones, see Fig. 9. The prototype presented can be seen to be a very promising candidate for the DTV antenna of a tablet-size terminal.

B. Design for a Typical-Size Terminal Including E-GSM

The purpose of this subsection is to demonstrate a simulated DTV antenna design for a typical-size terminal with realistic, non-ideal matching circuit implementation. The design takes place with the DTV antenna element dimensions of $d = l_{CCE} = 10$ mm and $h = 6.0$ mm see Fig. 17. The volume of the DTV coupling element used is 2.9 cm^3. According to Fig. 13 the margin to the realized gain specification is 2.8 dB when a 0.89 Ω additional resistor and ideal matching circuit components are used. The structure is considered to be a good compromise between the size and the performance. In addition, the effect of the E-GSM antenna mounted on the same chassis is taken into account in this design. The capacitive coupling element-based E-GSM antenna is designed according to [13]. The volume occupied by the E-GSM coupling element is 1.4 cm^3, see Fig. 17.

The realistic matching circuit components are chosen from the Murata's selection [35]. The inductors are from LQW18A and LQW15A series and the capacitors are from GJM15 and GJM03 series. The components are modeled with S parameters available in [36]. The E-GSM matching circuit is designed according to [33]. Both matching circuit topologies and the nominal values of the components are shown in Fig. 18. As discussed above, the relatively large-value series inductor $(L_m = 36\ \mathrm{nH})$ in the DTV matching circuit has relatively large series resistance [35] and thus, the additional resistor $(0.89\ \Omega)$ can be left out.

The simulated S parameters for DTV and E-GSM antennas are shown in Fig. 19. The realized gain of the DVB-H antenna and the total efficiency of the E-GSM antenna are shown in

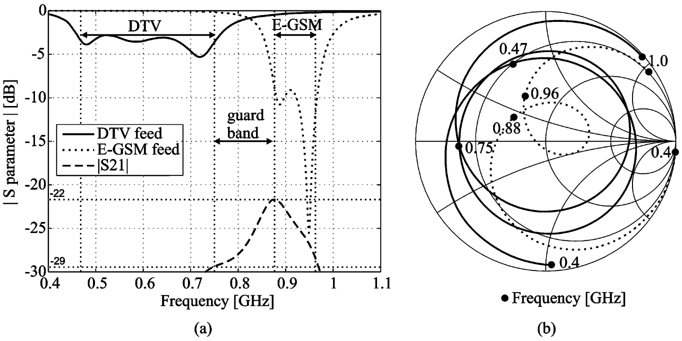

Fig. 19. Simulated S parameters of the typical-size terminal design (a) in the Cartesian coordinate system and (b) on the Smith chart.

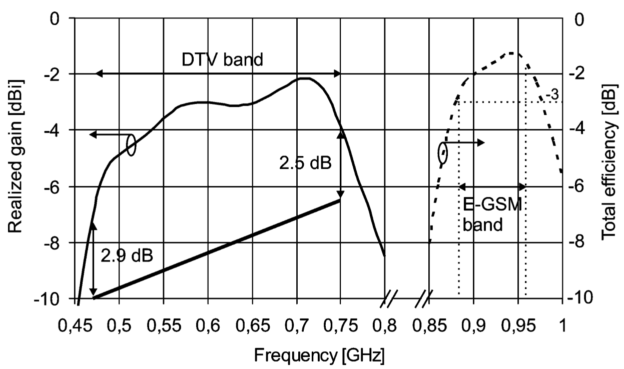

Fig. 20. Simulated realized gain of the DVB-H antenna and the simulated total efficiency of the E-GSM antenna of the typical-size terminal design.

Fig. 20. All the simulations are performed against 50 Ω reference impedance.

As can be seen, the return loss is at least 2.8 dB across the DTV band. The improvement of the matching level from 1.5 to 2.8 dB is due to the losses of the matching circuit components. On the Smith chart the reflection coefficient looks fairly optimal from the bandwidth point of view [22]. For the E-GSM antenna the return loss is at least 9 dB across the E-GSM band (0.88–0.96 GHz).

Below 0.5 GHz the realized gain of the DTV antenna drops several decibels compared to the ideal matching circuit implementation, e.g., in Fig. 8. Instead, at the upper DTV frequencies the realized gain remains roughly on the same level as in the ideal designs. This is due to two main reasons: First of all, at the lower DTV frequencies the radiation resistance is rather small (see Fig. 4) and thus the resistive losses in the matching circuit decrease the radiation efficiency more than at the upper DTV frequencies, where the radiation resistance is larger. Thus, it is very important that the losses of the matching circuit components are modeled realistically with the S parameters of real components. Secondly, the matching circuit is optimized in such a way that a part of the available margin at the lower DTV frequencies is transferred to the upper frequencies, where the margin to the realized gain is typically lower. Thus, the margin to the realized gain specification is optimized across the DTV band and it is at least 2.5 dB. Again, the total efficiency is 2 dB lower than the realized gain.

The total efficiency of the E-GSM antenna is higher than -3 dB (50%) across the E-GSM band. Relatively high losses are caused also here by the matching circuit components. The total efficiency can be improved using components with lower losses in the matching circuit.

Because the chassis is the main radiator below 1 GHz, both antenna elements mounted on the chassis are electromagnetically strongly coupled to the lowest order wavemode of the chassis and thus the elements are also strongly coupled to each other through the chassis [12]. Since the antennas are matched at different frequency bands, the matching circuits operate also as band pass filters [14]. Because the transition from the pass band to the reject band cannot be very sharp, the minimum isolations between the antennas are only 29 and 22 dB at the DTV and E-GSM bands, respectively, see Fig. 19. Thus, if the E-GSM transmits at its maximum power level, i.e., 33 dBm, 11 dBm is coupled to the DTV receiver and the operation of DTV would be blocked [23], [24]. Therefore, the matching circuits do not provide enough isolation and in order to make the simultaneous operation possible more isolation is needed. According to [24] 61 dB isolation between DTV and E-GSM at the E-GSM band is required, the next subsection is devoted for that issue. The isolations between DTV and the other transmitting systems (such as GSM1800, UMTS, Bluetooth and WLAN) are not problematic since the corresponding frequency bands are far enough from DTV band [14].

C. Improvement of the Isolation Between DTV and E-GSM

Basically there are two alternative ways how to improve the isolation between DTV and E-GSM. Firstly, one can use circuit technology such as filters, or secondly, one can increase the electromagnetic isolation, concept introduced in [39], by modifying the 3D antenna structure.

The first way exploiting an available GSM reject filter in the input of the DTV receiver is an easy solution from the antenna point of view. An alternative approach has been introduced in [40] where a combined matching and filtering circuit comprises a strong GSM trap which attenuates the GSM signals by more than 40 dB. The circuit-based solutions seem feasible but, however, they always introduce a certain insertion loss also for the DTV signals and thus they decrease the efficiency of the DTV antenna.

Using the second way the antenna structure might be modified in such a way that the electromagnetic isolation, which does not depend on the matching, is maximized. The current placing of the DTV and E-GSM antenna elements in the opposite ends of the chassis in Fig. 17 is not obviously optimal from the electromagnetic isolation point of view since both elements couple strongly to the same wavemode created by the longitudinal currents of the chassis. However, the current placing in the typical-size terminal is chosen in order to get enough impedance bandwidth for both systems. By reshaping the elements they might be placed in the corners of the same end of the chassis and hence the electromagnetic isolation might increase because the principal current paths for both elements are diagonally from corner to opposite corner and thus have some diversity. Other solutions might include, e.g., quarter-wavelength wavetraps [41] but it is possible that they are physically too large at the lower UHF frequency band to be placed inside typical-size terminals, or neutralization line which might produce opposite coupling as proposed in [42].

The best solution might be to somehow maximize the electromagnetic isolation according to the boundary conditions of a given mobile environment, and then provide the rest of the required isolation with the help of the circuit technology. However, finding an optimal way to fulfill the isolation requirement requires further work.

VI. Discussion

The capacitive coupling element-based broadband DTV antenna structures offer certain significant advantages. Firstly, the design of the studied antenna concept is relatively straightforward because the antenna matching is created with the matching circuit and thus the type, location and shape of the antenna element can be optimized according to the given boundary conditions of the mobile terminal environment. Secondly, the broadband antenna structures provide rather simple implementation compared e.g., to electrically tunable antennas based on a semiconductor component. Thirdly, the passive implementation of the broadband antenna structures does not cause any distortion problems when using the cellular system transmitters simultaneously with DTV.

The implementation of broadband DTV antennas is a compromise between the size and the performance of the antenna. In this work a 3-dB margin to the realized gain specification was reserved for the implementation losses introduced in commercial products. When applying the 3-dB margin to the specification it was shown with electromagnetic simulations that relatively high capacitive coupling elements are required especially in today's typical-size terminals. In order to make the antenna elements thinner without increasing the total volume of the elements, the performance of the antenna has to be slightly sacrificed.

The results of the paper also show a few apparent challenges of the studied antenna concept. Firstly, the efficiency of broadband DTV antennas is not as good as usually in mobile terminal transceiver antennas of cellular systems. However, the lower efficiency level can be accepted because DTV is a receive-only-system. Secondly, the isolation between the DTV and E-GSM systems of the simulated design for a typical-size terminal is not high enough to fulfill the specification given in [24] and thus additional filtering between DTV and E-GSM radios is required. Thirdly, the chosen matching circuit topology and the used components affect relatively much the performance of the antenna through the losses especially at the low DTV frequencies and the variation of the component values.

In this paper a traditional way to evaluate mobile terminal antennas has been used: the performance is calculated against 50 ohms reference impedance in terms of the total efficiency or realized gain. This approach is especially useful in academic publications since the comparison between different antennas is thereby rather easy to perform. However, a different approach is available for receive-only antennas. As well known, for receiving systems the sufficient signal-to-noise ratio (SNR) is the fundamental requirement. The SNR of a highly capacitive unmatched antenna and a given high-impedance low noise amplifier can be optimized to implement a broadband SNR response, as demonstrated in [43], [44]. The design principles of low noise amplifiers (LNA), introduced e.g., in [16], can be used by considering the input of the antenna as the source of the LNA. The method might also be useful for evaluating a CCE-based broadband DTV antenna and low noise amplifier receiver system. In addition, the method might also provide a new approach how to optimize the DTV antenna structure.

VII. Conclusions

The implementation and design of broadband internal digital television antennas for handheld multisystem mobile terminals is studied systematically. Since the volume that can be reserved for antennas inside a mobile terminal is very restricted, the design principle of (receive-only) digital television antennas is to sacrifice the total efficiency of the antenna to a level which is just enough to ensure a sufficient signal-to-noise ratio and that way make the size of the antenna element sufficiently small. Thus, the lower limit for the size of broadband capacitive coupling element-based antenna structures inside terminals of different size was studied with electromagnetic simulations. The results indicate that DTV antenna elements for today's small and typical size monoblock terminals are relatively high and large when reserving a 3-dB margin for the implementation losses of commercial terminals. It was shown that the height of a DTV antenna element can be made clearly smaller by sacrificing only slightly the 3-dB margin. On the other hand, for terminals whose chassis is longer than in typical size terminals, such as in tablet-size terminals, the height of a DTV antenna element can be significantly smaller. The derived results are demonstrated both with a prototype and a simulated design. The results show that the studied antenna concept is a promising candidate for broadband digital television antennas in mobile terminals. The results also increase understanding on the implementation of antennas based on the radiation of the finite ground plane, so, the presented methods could also be applied at other frequencies for other systems, e.g., for the spectrum sensing antenna of the cognitive radio [45].

Acknowledgment

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