Coexistence of a fixed communication system with an existing personal communication network

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Abstract

This paper deals with the proposal and performance evaluation of a Fixed Microwave Communication (FMC) system that shares the same bandwidth with a wireless personal communication system. In particular we focus here on the case of an existing UMTS network. The main application of the FMC system is for wireless connections between remote base stations and a core network access point in a dense urban environment, in particular when a wired connection is too expensive for service providers due to its implementation cost and lack of frequency spectrum. The mutual interference effects between the FMC and the existing UMTS systems are investigated by focusing on typical application scenarios. A receiving scheme where interference signals are firstly, detected and then canceled from the other received signals is proposed. The performance for the two interfering systems has been evaluated in terms of bit error rate by means of analytical approach and computer simulations. The obtained results clearly demonstrate the effectiveness of the proposed method, allowing an efficient bandwidth sharing between a FMC system and an existing UMTS network.

I. INTRODUCTION

In the last decade, wireless mobile communications were developed more extensively than other communication systems. Initially, all the carried traffic was voice, but at present the share of internet connections and services has gained a grow interest. This has lead to the need of a complex network architecture capable of supporting connections with different Quality of Service (QoS) constraints and integrating circuit-switched or packet-switched capabilities. In particular, the recent proposed UMTS network architecture is formed by two main parts, namely the UMTS Terrestrial Radio Access Network (UTRAN) that deals with all radio-related functionalities and the Core Network which is devoted to switching and routing calls and data connections to external networks. Moreover, in the UTRAN architecture one or more remote NodeBs, i.e., base stations, are planned and devoted to perform basic air-interface operations in a cell [1]. These are connected to a Radio Network Controller (RNC) usually through a fixed network. The RNC basic operations include congestion control, admission control and code allocation for a new wireless connection in the cells associated to the controlled NodeBs.

This paper investigates the possibility of implementing a wireless NodeB–RNC communication (Fig. 1) by means of an FMC system that shares the same bandwidth as that of the UMTS network. The coexistence of a personal communication network with a fixed service microwave system was discussed in the literature in [2]–[4] but still represents an open issue. In particular, in [2]–[4], the feasibility of adding a personal communication network to an existing fixed-service microwave network is analyzed. The performance of the Personal Communication Network (PCN) users under the influence of the fixed microwave system and the influence of PCN users on the fixed microwave system was investigated in order to highlight that spectrum sharing between these two systems is possible. Differently from [2]–[4], this paper focuses on the feasibility of a spectrum sharing between a high bit-rate fixed microwave system and the UMTS system in a dense urban environment using an improved cancellation scheme at NodeB where FMC transmission is detected and then canceled to mobile users communications.

The performance of the two systems is given in terms of bit error rate (BER). A suitable channel parameters acquisition scheme for the FMC system is then analyzed. The results shown here clearly demonstrate that the full bandwidth sharing between the proposed fixed microwave system and the UMTS system is feasible and represents a suitable solution to guarantee a wireless network interconnection between a NodeB and the core network in an urban environment where the implementation of a classical wired connection may be difficult or too expensive.

II. SYSTEM MODEL

Being the proposed FMC used to establish a high-data rate interconnection between remote NodeBs and the core network (i.e., NodeB–RNC), we have assumed for the FMC case to use a 16-QAM modulation scheme with a rectangular signal space constellation [5]. The baseband complex envelop of the transmitted signal results to be:

$$s_{FMC}(t) = \sum_{n=0}^{\infty} (A_{n,I} + jA_{n,Q})g(t - nT)$$

where $A_{n,I}$ and $A_{n,Q}$ are the information-bearing signal amplitudes of the quadrature components and $g(t)$ the signal pulse shaping assumed of the raised-cosine type with roll-off factor equal to 0.2. The bit rate for the FMC system has been assumed to be equal to
Fiber

Cellular and PDA users

Fig. 1. A possible scenario with coexistence of fixed and mobile communications.

25.36 Mb/s, corresponding, in case the of the 16-QAM modulation, to 3.84 Msymbols/s, i.e., the chip rate for the UMTS system sharing the same bandwidth, so that \( T = T_c \) where \( T_c \) is the UMTS chip time. For the FMC system under consideration, we have assumed that the transmitting and receiving antennas are placed on the top of high towers or buildings and are directional. However, due to the dense urban environment where this system is supposed to operate, the communication channel was assumed as a multipath quasi-static channel with \( L_1 \) paths, characterized by the impulse response given by:

\[
h_{\text{FMC}}(t) = \delta(t) e^{j\phi_{\text{FMC},1}} + \sum_{i=2}^{L_1} \alpha_{\text{FMC},i} \delta(t - \tau_{\text{FMC},i}) e^{j\phi_{\text{FMC},i}}
\]  

(2)

where \( \alpha_{\text{FMC},i} \) is the attenuation coefficient, \( \phi_{\text{FMC},i} \) the phase and \( \tau_{\text{FMC},i} \) the delay value for \( i \)-th path. We assume that \( |\tau_{\text{FMC},i} - \tau_{\text{FMC},j}| = nT \) \( \forall i \neq j \) and \( n = 1, 2, \ldots \) where \( \tau_{\text{FMC},i} \) and \( \tau_{\text{FMC},j} \) are the time delays introduced by the \( i \)-th path and the \( j \)-th path, respectively. From (1) and (2) it follows that the received signal, neglecting the effect of the UMTS system and additive white Gaussian noise (AWGN), has the form:

\[
\tilde{r}_{\text{FMC}}(t) = s_{\text{FMC}}(t) e^{j\phi_{\text{FMC},1}} + \sum_{i=2}^{L_1} \alpha_{\text{FMC},i} s_{\text{FMC}}(t - \tau_{\text{FMC},i}) e^{j\phi_{\text{FMC},i}}
\]  

(3)

Let us now introduce the UMTS system in order to evaluate the mutual interference of FMC system and mobiles.

According to the UMTS uplink standard [1], [6], [7], the baseband complex envelop of the transmitted signal for the \( k \)-th user with data bit rate \( R_k \), can be written as:

\[
s_{k} = \sum_{n=0}^{\infty} \sqrt{2P_k} \left[ \sum_{s=1}^{256/SF_k} d_{n,s}^k + \sum_{s=1}^{256/SF_k} c_{n,s}^k \right] O_{u}(t - nT_0 - sT_k) + j\beta c_n^k O_0(t - nT_0) S_{cl}(t - nT_0)
\]  

(4)

where:

- \( SF_k \) is the spreading factor of user \( k \) that varies from 4 to 256;
- \( d_{n,s}^k \) and \( c_{n,s}^k \) are the \((n \cdot 256 + s)\)-th data bit and the \( n \)-th control bit of the \( k \)-th user, respectively;
- \( O_u(t) \) is the OVSF (Orthogonal Variable Spreading Factor) code formed by rectangular pulses (chips) of duration \( T_c \) and with a random amplitude ±1; the ID number \( n \) is selected according to the user spreading factor (SF\( k \)). The code length is SF\( k \) for the in-phase bit stream and 256 for the quadrature bit stream.
TABLE I

<table>
<thead>
<tr>
<th>Tap number</th>
<th>Relative time (µs)</th>
<th>Average relative power (dB)</th>
<th>Doppler spectrum</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>0.0</td>
<td>0.0</td>
<td>CLASS</td>
</tr>
<tr>
<td>2</td>
<td>0.25</td>
<td>-2.0</td>
<td>CLASS</td>
</tr>
<tr>
<td>3</td>
<td>0.5</td>
<td>-1.0</td>
<td>CLASS</td>
</tr>
<tr>
<td>4</td>
<td>0.75</td>
<td>-9.0</td>
<td>CLASS</td>
</tr>
</tbody>
</table>

- $S_k(t)$ is the complex scrambling code of $k$-th user; it is formed by 256 root raised-cosine pulses (chips) with roll-off factor $0.22$ of duration $T_c$ and amplitude $\pm 1$;
- $T_{bk}$ is the data bit duration of user $k$ ($T_{bk} = 1/R_k$), $T_0$ is the control bit duration ($T_0 = 1/(15\text{kb/s})$);
- $P_k$ is the transmission power of the $k$-th user and $\beta$ is the power imbalance between data and control streams.

In a system where $K$ is the number of the simultaneous active users and $L_2$ is the number of resolved multipath components, the received signal (without considering the influence the FMC system and AWGN) can be defined as:

$$r_m(t) = \sum_{k=1}^{L_2} \sum_{l=1}^{L} a_{m,l}^k s_{m,l}^k (t - \tau_{m,l}^k) e^{j\phi_{m,l}^k}$$  \hspace{1cm} (5)

where $a_{m,l}^k$, $\phi_{m,l}^k$, $\tau_{m,l}^k$ are attenuation, phase, and delay, respectively, of the $l$-th replica of the $k$-th transmitted signal. These are independent random variables whose statistical distributions are Rayleigh for $a_{m,l}^k$, uniform in $[0, 2\pi)$ for $\phi_{m,l}^k$, and uniform in $[0, T_0]$ for $\tau_{m,l}^k$. The parameters ($a_{m,l}^k$, $\phi_{m,l}^k$, $\tau_{m,l}^k$) and the spreading sequence of all mobile users are assumed to be known at the receiving end. The channel considered herein is characterized by 4 resolvable paths according to the channel model [8] shown in Tab. I.

In the proposed system (Fig. 2), at the receiving end, the FMC and the UMTS mobile signals are collected together. The FMC antenna is supposed to be directive and the mobiles interference is received by secondary lobes with an attenuation factor $\gamma$, that takes into account that not all the $k$ user signals are captured by the FMC antenna. Hence, the total received signal at FMC antenna results to be:

$$r_{\text{FMC}}(t) = \tilde{r}_{\text{FMC}}(t) + \gamma \tilde{r}_m(t) + n_1(t)$$  \hspace{1cm} (6)

On the other hand, the omnidirectional antenna, used for the mobiles, capture all the FMC signal:

$$r_m(t) = \tilde{r}_m(t) + \tilde{r}_{\text{FMC}}(t) + n_2(t)$$  \hspace{1cm} (7)

In (6) and (7), the terms $n_1(t)$ and $n_2(t)$ represent two independent AWGN contributions introduced by the appropriate receiving schemes, with zero mean and two side power spectral density equal to $N_{0_1}/2$ and $N_{0_2}/2$, respectively. In Section V it is shown how interference from FMC system on mobile communications is more problematic than mobile communications interference toward the FMC system. Hence, in order to reduce this drawback a suitable cancellation scheme has been considered.

The FMC receiver is based on a decision feedback equalizer (DFE) that performs interference cancellation to avoid BER degradations due to the influence of intersymbols interference (ISI) caused by FMC signals replicas received. Hence, the signal at the input of the FMC symbol detector is:

$$v(t) = \tau_{\text{FMC}}(t) - \sum_{i=2}^{L_1} \alpha_{\text{FMC},i} \cdot \left\{ \sum_{n=L_i}^{\infty} \left( \hat{A}_{(n-N_i)} + j\hat{A}_{(n-N_i)} \right) g(t - i\tau_{\text{FMC},i} - nT) e^{j\phi_{\text{FMC},i}} \right\}$$  \hspace{1cm} (8)

where $\hat{A}_{n1}$ and $\hat{A}_{n2}$ are the information-bearing signal amplitudes of the quadrature carriers related to the $i$-th detected symbol and $N_i = \tau_{\text{FMC},i}/T$.

The detected symbols are also used to restore the total FMC signal and to remove its interference from $r_m(t)$ [1], [9]. Complex envelop of the restored signal is:

$$\tilde{r}_{\text{FMC}}(t) = \sum_{i=1}^{L_1} \alpha_{\text{FMC},i} e^{j\phi_{\text{FMC},i}} \cdot \sum_{n=0}^{\infty} (\hat{A}_{nt} + j\hat{A}_{nt}) g(t - i\tau_{\text{FMC},i} - nT)$$  \hspace{1cm} (9)

Hence, the signal to be processed by a bank of Rake receivers\(^1\), one for each active mobile user, in order to produce symbol detection is:

$$\tilde{r}_m(t) = r_m(t) - \tilde{r}_{\text{FMC}}(t) = \tilde{r}_m(t) + n_2(t) + \epsilon(t)$$  \hspace{1cm} (10)

where $\epsilon(t)$ is the cancellation error term.

\(^1\)The performance of the symbol detection process for mobile users may be improved by resorting to more efficient detection scheme as that proposed in [10].
III. CHANNEL PARAMETER ACQUISITION AND TRACKING

In order to implement coherent symbol detection at the FMC system receiving end, efficient channel parameter acquisition and tracking schemes have to be implemented. Being the communication channel a quasi-static multipath channel, the acquisition operation for the channel parameters (i.e., amplitude $\alpha_{\text{FMC},i}$, delay $\tau_{\text{FMC},i}$, and phase $\phi_{\text{FMC},i}$) for each path can be carried out on a long temporary scale because of their very slow variation in time. As a consequence, this operation can be performed at the set up of a communication session or whenever required by the data link layer as a reaction to an anomalous working situation, e.g., the detection of a high error rate condition. During the acquisition phase, we suppose to transmit a PN signal of length $M$, known at the receiver, (pilot signal) in order to perform the channel parameter estimation, defined as:

$$\tilde{g}_{E}(t) = A g_{E}(t) = \sum_{m=1}^{M} A c_{m}^{(E)} g(t - nT)$$  \hspace{1cm} (11)

where $A$ denotes the power amplitude of the pilot signal, and $c_{m}^{(E)} (= \pm 1)$ are the binary chip forming the PN sequence. It is a straightforward matter to see that the received signal (baseband) is:

$$\tilde{r}_{\text{FMC}}(t) = \sum_{i=1}^{L_1} \alpha_{\text{FMC},i} \tilde{g}_{E}(t - \tau_{\text{FMC},i}) e^{j\phi_{\text{FMC},i}} + \gamma r_{m}(t) + n_{1}(t).$$  \hspace{1cm} (12)

The digital detection scheme considered in this section allows the channel variations to be acquired by evaluating the sliding correlation of the received signal $\tilde{r}_{\text{FMC}}(t)$ with a locally generated sequence $g_{E}(t)$. By taking into account the assumption of a very slow fading (i.e., the channel parameters remain constant over several consecutive bit intervals), we have:

$$c(\tau) = \int_{\tau}^{\tau + MT} \tilde{r}_{\text{FMC}}(t)g_{E}(t - \tau)dt = \sum_{i=1}^{L_1} \left[ \alpha_{\text{FMC},i} A e^{j\phi_{\text{FMC},i}} \xi(\tau - \tau_{\text{FMC},i}) \right] + \sigma(\tau)$$  \hspace{1cm} (13)

where:

- $\tau \in \left[0, T_{\text{search}}\right]$, with $T_{\text{search}}$ represents the search window width;
- $\xi(x)$ is equal to 1 if $x = 0$, or to zero, otherwise;
- $\sum_{i=1}^{L_1} \alpha_{\text{FMC},i} A e^{j\phi_{\text{FMC},i}} \xi(\tau - \tau_{\text{FMC},i})$ is the contribution of the auto-correlation between $g_{E}(t)$ and every received replica of $\tilde{g}_{E}(t)$ when $\tau = \tau_{\text{FMC},i}$;


- \( \sigma(\tau) \) is the contribution due to the multiple access interference (i.e., term \( \tilde{r}_m(t) \) in (6)), AWGN term (i.e., \( n_1(t) \) in (6)) and self-interference.

In order to describe the mode of operation of the channel parameters estimation method, we consider an ideal condition by assuming \( \sigma(\tau) \) as negligible. According to this, the squared module of \( c(\tau) \) results to be:

\[
D(\tau) = (\text{Re}\{c(\tau)\})^2 + (\text{Im}\{c(\tau)\})^2 = \sum_{i=1}^{L_1} \alpha_{\text{FMC},i}^2 \alpha^2(\tau - \tau_{\text{FMC},i}).
\] (14)

In the classical sliding correlation approach the delays associated with the \( L_1 \) paths of interest can be derived by searching for the \( L_1 \) values of \( \tau_j \), with \( j = 1, \ldots, L_1 \) for which \( D(\hat{\tau}_{\text{FMC},i}) \) assumes maximum values. Hence, the channel path attenuation and phase are derived as:

\[
\hat{\alpha}_{\text{FMC},i} = \sqrt{D(\hat{\tau}_{\text{FMC},i})}, \quad \text{for } i = 1, 2, \ldots, L_1
\]

\[
\hat{\phi}_{\text{FMC},i} = \tan^{-1}\left[\frac{\text{Im}\{c(\hat{\tau}_{\text{FMC},i})\}}{\text{Re}\{c(\hat{\tau}_{\text{FMC},i})\}}\right], \quad \text{for } i = 1, 2, \ldots, L_1
\] (16)

Although the scheme is extremely simple to implement, the channel estimates may be inaccurate owing to the influence of ISI. In order to increase the channel parameters estimation accuracy, a successive estimation/cancellation method has been used [11]. At the first step, the greatest \( D(\tau_j) \) values with the correlation interval \([0, kT]\) is considered. Channel parameters are derived accordingly to (15) and (16). Hence, the associated pilot signal component is regenerated and removed from the received signal. This procedure is started over again until the channel parameters for the \( L_1 \) path have been derived.

During the tracking (or steady-state) mode of operation, following the acquisition (or startup) phase, channel parameters estimation is limited to the carrier phase and symbol timing, due to their higher sensitivity to the wireless channel variations and receiver oscillator drift. Under the assumption that carrier phase and symbol timing vary slowly on two consecutive symbol intervals, decision-direct carrier phase and symbol timing estimators used in the feedback mode were considered, i.e., the carrier phase and symbol timing estimations derived on the basis of the actual received symbol knowledge are used to detect the next received symbol. Moreover, it is important to point out that carrier phase and bit timing estimates are derived after removing ISI from the received signal due to past received signals. For the symbol timing estimation, a modified version of the well-known Early-Late Gate scheme suitable for 16-QAM signal was considered. The block diagram of the Early-Late Gate synchronizer proposed for the FMC system under consideration is sketched in Fig. 3.

As in the classical Early-Late Gate circuit, the output of the matched filter, representing the autocorrelation of the received signal is sampled early at \( t = T - \delta \) and late at \( t = T + \delta \) with respect to the time instant \( t = T \) of the peak of the correlation function [5]. The timing estimation accuracy is dependent on \( \delta \) and hence on the number of samples per symbol. After an investigation of this point by computer simulations, the most suitable number of samples per symbol selected on the basis of the best trade-off between implementation complexity and timing estimation accuracy, has been resulted equal to 11. A decision-direct implementation is considered so that the influence of the transmitted symbol is removed by multiplying the outputs of the correlator by the inverse of the information-bearing signals amplitude associated with the in-phase detected symbol, \( \hat{A}_k \). The same is possible by focusing on the quadrature channel. The error function for the synchronization process is then obtained as the difference between the two early and late samples. Hence, after removing the high-frequency components by means of the loop filter, the error signal is used to drive the Voltage Controlled Oscillator (VCO) as shown in Fig. 3.

Also for the carrier phase tracking, a decision direct approach based on the Maximum Likelihood (ML) criteria [5] was considered (Fig. 4). The basic assumption was that of assuming the carrier phase varies very slowly on two symbol intervals.

In order to illustrate the mode of operation of the considered carrier phase estimation scheme we consider as negligible the AWGN contribution and a perfect IC, so that the signal is not affected by ISI. Based on the above assumptions the signal at the input of the carrier phase estimation circuit (Fig. 4) is:

\[
y(t) = A_{k,I}g(t) \cos(2\pi f_c t + \phi) - A_{k,Q}g(t) \sin(2\pi f_c t + \phi).
\] (17)

This signal is firstly multiplied by the quadrature carriers \( \cos(2\pi f_c t + \phi) \) and \( \sin(2\pi f_c t + \phi) \), derived by the VCO by assuming for the carrier phase estimation \( \phi \) the value obtained on the previous symbol interval. The multiplication of \( y(t) \) by \( \cos(2\pi f_c t + \phi) \) followed by match filtering to \( g(t) \) and sampling at the symbol rate yields the in-phase term to:

\[
y_{k,I} = \frac{1}{2} A_{k,I} \cos(\phi - \hat{\phi}) - \frac{1}{2} A_{k,Q} \sin(\phi - \hat{\phi}) + \text{double-frequency terms}
\] (18)

Likewise, multiplication of \( y(t) \) by \( \sin(2\pi f_c t + \phi) \) followed by match filtering to \( g(t) \) and sampling at the symbol rate yields the quadrature term to:

\[
y_{k,Q} = \frac{1}{2} A_{k,Q} \cos(\phi - \hat{\phi}) + \frac{1}{2} A_{k,I} \sin(\phi - \hat{\phi}) + \text{double-frequency terms}
\] (19)
Fig. 3. Early-Late Gate circuit for 16-QAM signals.

Fig. 4. Phase estimation circuit for 16-QAM signals.
Hence, terms $y_{k,I}$ and $y_{k,Q}$ are combined together to form the error signal

$$e_k = y_{k,I} \frac{2\hat{A}_{k,I}}{A_{k,I}^2 + \sigma_{n,I}^2} - y_{k,Q} \frac{2\hat{A}_{k,Q}}{A_{k,Q}^2 + \sigma_{n,Q}^2}$$

(20)

where $\hat{A}_{k,I}$ and $\hat{A}_{k,Q}$ are the estimations of the in-phase and quadrature components of the received 16-QAM signal on the symbol interval of interest. Thus, in the absence of decision errors we have:

$$\hat{A}_{k,I} = A_{k,I}, \quad \hat{A}_{k,Q} = A_{k,Q}$$

(21)

Under this assumption, the error signal $e_k$ at the input of the loop filter results to be:

$$e_k = \sin(\phi - \hat{\phi}) + \text{double-frequency terms}.$$ 

(22)

The loop filter is a low-pass type, so that it rejects the double-frequency terms in $e_k$. The desired term is $\sin(\phi - \hat{\phi})$ that, approximated by $(\phi - \hat{\phi})$ gives the phase error term to drive the VCO, hence obtaining an updating of the carrier phase estimation to be used on the next symbol time interval.

IV. INTERFERENCE BETWEEN FIXED AND MOBILE NETWORKS

The main drawback for the considered FMC system is due to the spectrum sharing with the assumed UMTS network that produces a mutual interference and hence strongly reduce the performance. In this work we have focused our attention to the interference from UMTS reverse link to FMC system and interference from FMC system to UMTS NodeB. In this section these types of interference will be exploited in a theoretical way as described below.

A. Interference from UMTS reverse link to FMC system

This type of interference occurs when signal transmitted by User Equipments (UEs) of the UMTS network are received by FMC antenna jointly with the signal of interest (fixed communication). In this case, it is possible to derive a theoretical estimation of total interference power as outlined below.

Let us assume a static environment, i.e., FMC link without users interference. It follows that the Signal-to-Noise Ratio (SNR) results to be:

$$\text{SNR}_{\text{FMC}} = \frac{P_{\text{FMC}}}{\sigma_n^2}$$

(23)

where $P_{\text{FMC}}$ is the received power for the FMC signal at the receiver and $\sigma_n^2$ is the variance of the AWGN introduced by the FMC communication channel. Conversely, in the presence of UMTS users’ interference, we have to evaluate the Signal-to-Interference-and-Noise Ratio (SINR), defined as:

$$\text{SINR}_{\text{FMC}} = \frac{P_{\text{FMC}}}{\sigma_n^2 + P_t}$$

(24)

where $P_t$ is the UMTS users total interference power at the FMC receiving end. Without loss of generality, we will assume in the following that:

$$\sigma_{n_1}^2 = \sigma_{n_2}^2 = \sigma_n^2 \quad (N_{0_1} = N_{0_2} = N_0)$$

(25)

where $\sigma_{n_2}^2$ is the variance at the UMTS NodeB receiving antenna. The level of the interference contribution due to reverse UMTS communication link depends on the $E_b/N_0$ value needed to guarantee a target BER to each specific UMTS mobile signal, where $E_b$ is the energy per bit and $N_0$ is the one-sided noise spectral density. Under the assumption that each UMTS user may have only one ongoing service, we have for the user $k$:

$$\left( \frac{E_b}{N_0} \right)_k = \frac{P_k T_{b_k}}{\sigma_n^2 2 T_c} = \frac{P_k}{2 \sigma_n^2} \text{SF}_k$$

(26)

where $P_k$ is the power of the signal at FMC antenna for a generic mobile user, $T_{b_k}$ is the bit time duration for the user $k$, $T_c$ is the chip time and $\text{SF}_k = \frac{T_c}{T_{b_k}}$ is the processing gain. From (26), we have also:

$$P_k = \left( \frac{E_b}{N_0} \right)_k \frac{2 \sigma_n^2}{\text{SF}_k}$$

(27)

By taking into account that due to highly directive receiver antenna, we have that just $\gamma_{i}$ of the total user $k$ received power interferes with the FMC, and just $\gamma_{r}$ of the received power come into receiver for attenuation due to secondary lobes ($\gamma = \gamma_{i} \cdot \gamma_{r}$). Hence, we have:

$$P_t = \frac{\sigma_n^2}{\gamma_i \gamma_r} \sum_{k=1}^{N_u} \left( \frac{E_b}{N_0} \right)_k \frac{2}{\text{SF}_k}$$

(28)
where $N_u$ is the total number of interfering users. In particular, for the case of $N_u$ users with a same transmission rate (i.e., $SF_k = SF \forall k$), (28) becomes:

$$P_t = \frac{N_u E_b 2\sigma_n^2}{\gamma_i \gamma_r N_0 \text{SF}}$$  

(29)

Hence, it follows that:

$$\text{SINR}_{\text{FMC}} = \frac{P_{\text{FMC}}}{\sigma_n^2 + P_t}$$

$$= \frac{P_{\text{FMC}}}{\sigma_n^2 + \frac{\sigma_n^2}{\gamma_i \gamma_r} \sum_{k=1}^{N_u} \left( \frac{E_b}{N_0} \right) \frac{2}{SF_k}}$$

(30)

$$= \frac{\text{SNR}_{\text{FMC}}}{1 + \frac{1}{\gamma_i \gamma_r} \sum_{k=1}^{N_u} \left( \frac{E_b}{N_0} \right) \frac{2}{SF_k}}$$

B. Interference from FMC to UMTS NodeB

In order to investigate the interference effects due to FMC on an UMTS NodeB receiver, we have considered that both UMTS NodeB and FMC antenna are located at the same site. Let us focus on a particular UMTS user whose received power is $P_m$. By neglecting the contribution of the other interfering UMTS users, (i.e., the Multiple Access Interference - MAI), on the SINR, at the UMTS NodeB receiver we have:

$$\text{SINR}_m = \frac{P_m}{\sigma_n^2 + P_t}$$  

(31)

where parameter $\sigma_n^2$ is related to AWGN power at the input of the UMTS NodeB receiver and $P_t$ is the power of the interfering signal due to FMC.

Under the assumption of a constant power spectral density for the FMC signal in the channel bandwidth $B_c$ we have:

$$P_i = \frac{P_{\text{FMC}}}{B_c} B_m = \frac{P_{\text{FMC}}}{\text{SF}_m}$$

(32)

where $B_m$ is the one-sided bandwidth of the mobile signal under consideration and $\text{SF}_m$ is the associated process gain.

Furthermore, by recalling that $\text{SNR}_{\text{FMC}}$ is defined as:

$$\text{SNR}_{\text{FMC}} = \frac{P_{\text{FMC}}}{\sigma_n^2},$$

(33)

we have:

$$P_{\text{FMC}} = \sigma_n^2 \text{SNR}_{\text{FMC}}.$$  

(34)

From (33) it follows that (32) becomes:

$$P_i = \frac{\sigma_n^2 \text{SNR}_{\text{FMC}}}{\text{SF}_m}$$

(35)

Under the assumption of an equal power spectral density of the AWGN (see (25)) at the input of both FMC and UMTS NodeB receiver, it is straightforward to verify that:

$$P_i = \sigma_n^2 \text{SNR}_{\text{FMC}}$$

(36)

Finally, we have:

$$\text{SINR}_m = \frac{P_m}{\sigma_n^2 + \sigma_n^2 \text{SNR}_{\text{FMC}}}$$

$$= \frac{P_m}{\sigma_n^2 (1 + \text{SNR}_{\text{FMC}})}$$

$$= \frac{\text{SNR}_m}{1 + \text{SNR}_{\text{FMC}}}$$

(37)

Degradation (in dB) on the Bit Error Rate (BER) performance at the UMTS NodeB are given in Tab. II for different $\text{SNR}_{\text{FMC}}$ values. At this point, it is important to note that, in order to take into account the MAI effects due to simultaneous mobile user communications, parameter $\text{SINR}$ has to be rewritten as:

$$\text{SINR}_m = \frac{P_m}{\sigma_n^2 (1 + \text{SNR}_{\text{FMC}}) + \sigma_{\text{MAI}}^2}$$

(38)

where $\sigma_{\text{MAI}}^2$ is the MAI mean power $^2$.

$^2$Eq. (38) is written under the assumption of Gaussian Approximation (GA) for the MAI [12].
TABLE II
PERFORMANCES DEGRADATION DUE TO FMC COMMUNICATIONS.

<table>
<thead>
<tr>
<th>SNR_{FMC}(dB)</th>
<th>Degradation(dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>6</td>
<td>0.37</td>
</tr>
<tr>
<td>8</td>
<td>8.64</td>
</tr>
<tr>
<td>10</td>
<td>10.41</td>
</tr>
</tbody>
</table>

Fig. 5. Performance in terms of synchronization error for proposed timing recovery circuit.

V. NUMERICAL RESULTS

This section deals with the numerical results concerning the performance of the FMC under consideration. Fig. 5 shows the synchronization error, defined as difference in symbol time with respect to the correct timing, during the recovery phase. An observation period equal to 100 bit was considered for the case of synchronization errors equal to 4 samples (11 samples for each symbol-time were assumed). Fig. 5 highlights the good performance of proposed synchronization recovery circuit, in particular it is evident in this figure that a zero estimation error is achieved after 35 bit times. In Fig. 6 the performance of the FMC system is shown in terms of BER as a function of the SNR at the input of the receiver with channel parameters acquisition and tracking carried out according to the schemes outlined in the previous section. In the same figure, the BER performance obtained in the case of a perfect knowledge of the channel parameters at the receiving end (ideal case), i.e., without performing acquisition and tracking operations, and in the case of channel acquisition carried out according to the classical sliding window correlation approach, i.e., without performing ranking and successive cancellation of the detected signal replicas by using a DFE, are also reported for comparison purposes. The good behavior of the proposed channel parameters estimation approach is evident in this figure.

Performance degradations due to coexistence of the FMC system and an existing UMTS system in terms of BER have been derived by resorting to computer simulations. For UMTS mobile user communications, the propagation environment has been derived from [6], [7]: the number of resolvable path is assumed to be \( L_2 = 4 \). In performing the simulations we have assumed:
- QPSK modulation;
- Variable bit rate equal to 60kb/s and 960kb/s;
- Two coding levels: OVSF and scrambling sequences;
- Variable number of users and variable system load.

For the FMC system a two path environment has been considered (i.e., \( L_1 = 2 \)) with secondary path having an attenuation factor
\( a_{\text{FMC,2}} \) equal to -7 dB. Delay of the secondary path has been considered \( \tau_{\text{FMC,2}} = T \). We have also supposed that the value of \( \gamma = \gamma_1 \gamma_r \), i.e., the parameter that takes into account the attenuation of the UMTS mobile signals entering the FMC antenna, is 1/10 due to secondary lobes attenuation and due to a minor number of interfering users captured by the FMC antenna respect to the total transmitting UMTS mobile users inside the cell. Besides, performance are quite similar considering \( \gamma < 1/5 \).

In Fig. 7 the performance of the FMC system with 2 interfering UMTS users at 960kb/s is considered [1], in the case of a two-path channel and a DFE at the receiving end as outlined in Section II. The numerical results obtained via computer simulation are also compared with analytical predictions according to the approach outlined in Section IV-A. It is possible to note that the interfering users with \( E_b/N_0=12 \text{dB} \) do not significantly influence the performance of the considered FMC system. Moreover, a good agreement between simulation results and analytical predictions derived in Section IV-A is evident in the figure.

Fig. 8 shows the BER performance for the NodeB receiver. These results have been derived by assuming to have only one mobile user in order to neglect the MAI phenomenon at the receiving end. It is possible to note that the degradation introduced by the FMC signal is very close to that foreseen by means of the analytical approach outlined in Section IV-B and reported in Tab. II.

In Figs. 9 and 10 the performance of NodeB receiver is considered. In these figures parameters SNRF and SNRF\text{canc} denote the SNR for the FMC system at BS antenna with and without interference cancellation. The effects of the proposed cancellation scheme is considered for different values of SNR or number of active UMTS mobile users at receiving end. The results shown in Figs. 9 and 10 highlight that the performance degradation is strongly reduced by resorting to the proposed cancellation approach. It is also important to note that when the cancellation scheme is performed the performance are better for higher SNR of FMC system. This is due to fact that under high SNR, we have a more reliable estimation of transmitted symbol and so a more reliable cancellation of interfering signal. In Fig. 10 performance comparison is done for the case of a multirate environment where one UMTS user transmits at 960kb/s and 16 UMTS users transmit at 60kb/s; performance are evaluated for a target 60kb/s UMTS user. The performance advantages of the proposed scheme are evident in this figure.

Finally, the effects of the FMC channel parameter acquisition is considered for both FMC and UMTS NodeB performance. In Fig. 11 the performance for the FMC system with 8 interfering users at 60kb/s is shown by taking into account the effect of cancellation both in the case of a perfect knowledge of channel parameter and effective parameter estimation. The figure shows that considered channel parameter estimation method permits to have a slight performance degradation respect to the ideal channel parameter estimation.

In Fig. 12 the performance of an UMTS NodeB receiver in the case of 8 users, each one transmitting at 60kb/s, is considered. The good behavior of the proposed schemes is evident in the figure. In particular we have a low degradation with respect to the
Fig. 7. BER performance of the FMC system for the case of 2 UMTS interfering users transmitting at 960kb/s.

Fig. 8. BER Performance for the UMTS system (single user) with and without the interfering FMC signal.
Fig. 9. BER performance for the UMTS system with the interfering FMC signal in case of 2 mobile users transmitting at 960kb/s.

Fig. 10. BER performance in case of 17 multirate mobile users; user of interest transmits at 60kb/s.
Fig. 11. BER performance for the FMC system in the case of 8 mobile interfering users transmitting at 60kb/s.

case of the use of the IC approach under the assumption of an ideal knowledge of all the channel parameters of interest. As a minor remark, it is possible to say that, in the case of a channel parameter estimation, there is a trade-off in performance between an high SNRf, used for obtaining a better FMC detection, and a reduced efficiency of the cancellation process, due to not ideal parameters estimates; SNRf_est is referred to the case in which cancellation and channel parameter estimation are performed.

VI. CONCLUSIONS

In this paper the possibility of a spectrum sharing between an advanced FMC system and an UMTS system has been investigated. Both systems were assumed to operate in an urban environment. Proposed system consists in an advanced receiver that permits to cancel mutual interference between considered systems. A suitable acquisition channel parameter estimation schemes was also analyzed. The numerical results shown in this paper clearly demonstrate the good performance of the FMC system and the feasibility of spectrum sharing between this system and an existing UMTS system with the proposed cancellation scheme. We stress that this result is of particular interest in an urban environment for a fast implementation of a connection between remote UMTS NodeBs and the UMTS core network.

REFERENCES

Fig. 12. BER performance in case of 8 mobile users transmitting at 60kb/s.
