

ISTANBUL TECHNICAL UNIVERSITY ★ GRADUATE SCHOOL

**NEW RF ENERGY HARVESTING MODELS FOR
NEXT-GENERATION WIRELESS COMMUNICATION SYSTEMS**



Ph.D. THESIS

Mohammadreza BABAEI

Applied Informatics Department

Information and Communications Engineering Programme

NOVEMBER 2022

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**YENİ NESİL TELSİZ İLETİŞİM SİSTEMLERİ İÇİN
YENİ RF ENERJİ HASATLAMA MODELLERİ**

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To my dear and kind Parents, and Brothers



FOREWORD

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ABBREVIATIONS

AF	: Amplify-and-forward
AWGN	: Additive white Gaussian noise
BEP	: Bit error probability
BER	: Bit error rate
BPSK	: Binary phase shift keying
CDF	: Cumulative distribution function
CR	: Cognitive radio
CSI	: Channel state information
D	: Destination
DC	: Direct current
DBPSK	: Differential-BPSK
DF	: Decode-and-forward
DH	: Dual-hop
EH	: Energy harvesting
ER	: Energy receiver
FD	: Full-duplex
GRS	: General relay selection
HD	: Half-duplex
i.i.d	: Independent and identically distributed
IoT	: Internet of Things
IP	: Information processing
LoS	: Line of sight
MIMO	: Multiple-input multiple-output
MISO	: Multiple-input single-output
MRC	: Maximal-ratio-combining
NOMA	: Non-orthogonal multiple access
NL-EH	: Nonlinear-EH
OFDM	: Orthogonal frequency-division multiplexing
OFDMA	: Orthogonal frequency-division multiple access

PB	: Power beacon
PDF	: Probability distribution function
PR	: Primary receiver
PS	: Power-splitting
PT	: Primary transmitter
PU	: Primary users
R	: Relay
RF	: Radio frequency
RSSE	: Root-sum-of-squares error
r.v	: Random variables
S	: Source
SDF	: Selective-DF
SEP	: Symbol error probability
SNR	: Signal-to-noise-ratio
SR	: Secondary receiver
ST	: Secondary transmitter
SU	: Secondary users
SWIPT	: Simultaneous wireless information and power transfer
TS	: Time-switching
TW	: Two-way
UAV	: Unmanned aerial vehicle
WP	: Wireless powered
WPC	: Wireless-powered communication.
WPT	: Wireless power transfer

SYMBOLS

${}_pF_q(a, b; c; \cdot)$	: Hyper-geometric function
${}_1F_1(\cdot; \cdot; \cdot)$	: Kummer Confluent Hyper-geometric function
$\mathbf{K}_\nu(\cdot)$	: Modified Bessel function of the second kind
$G_{pq}^{mn} \left(\cdot \middle \begin{smallmatrix} a_p \\ b_q \end{smallmatrix} \right)$	: Meijer-G function
$\text{erfc}(\cdot)$	: Complementary error function
$\mathbf{Q}(\cdot)$	: Q function
$\Gamma(\cdot)$	: Gamma function
$\Gamma(\cdot, \cdot)$	: Upper Gamma function
$\gamma(\cdot, \cdot)$	: Lower Gamma function
$\mathbf{f}(\cdot)$	: Probability distribution function (PDF)
$\mathbf{F}(\cdot)$	: Cumulative distribution function (CDF)
ν	: The path loss factor
η	: Energy conversion coefficient ($0 < \eta \leq 1$)
$\mathbf{N}_0$	: Noise power spectral density



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NEW RF ENERGY HARVESTING MODELS FOR NEXT-GENERATION WIRELESS COMMUNICATION SYSTEMS

SUMMARY

Deployment of massive sensor and Internet of Things (IoT) devices in next-generation wireless communication systems reveals energy limitations as one of the challenges. Typically, IoT devices are powered by a battery, which restricts their capacity and working time. Energy harvesting (EH) has been regarded as a promising approach which can increase the life-time of a wireless communication system. In EH, energy is obtained by wind, solar, vibration, etc. Thus, the harvested energy is transformed into electricity and can be used by the desired nodes. However, the aforementioned traditional EH methods are not always available. Additionally, radio frequency (RF) EH has emerged as a key promising technique that enables wireless systems to harvest energy from the incoming signals in the environment. This energy is available as dedicated or ambient energy and can be harvested throughout the whole day. Hence, RF EH can be an effective alternative for empowering battery-free IoT devices, resulting in increased operational time. As RF EH methods for power-constrained nodes, simultaneous wireless information and power transfer (SWIPT) and wireless-powered communication (WPC) schemes are studied in the literature. The SWIPT scheme employs two EH receiver structures: power-splitting (PS) and time-switching (TS). The power-constraint node harvests power from the incoming signal energy in PS EH mode, with one portion of the signal power used for harvesting energy and the remainder for information processing. On the other hand, in the TS EH mode, EH and information processing (IP) are allocated to two non-overlapping time intervals, respectively. Furthermore, the amount of harvested energy is regarded as a linear or nonlinear (NL) function of the input power of the energy receiving (ER) node. The input and output powers of the EH circuit are directly proportional in the linear EH model. For the region with low input power, the NL EH model provides the same amount of energy as the linear model. However, the amount of harvested power saturates to a predefined threshold power at the high input power of the EH circuit.

In this thesis, three different RF EH system models are investigated and the closed-form analytical expressions for their performances are derived. Moreover, the performance of each considered system is investigated in terms of bit error probability (BEP) and outage probability. Besides, theoretical derivation results are provided for different system parameters and supported by the Monte-Carlo simulation results. Comprehensive insights into the studied systems are provided, which broaden the view of engineers toward the EH system designs.

The first section investigates the bit error rate (BER) performance of a full-duplex (FD) overlay cognitive radio (CR) network with linear/NL EH capability. The studied overlay CR network comprises of a primary transmitter/receiver (PT/PR) pair, a secondary transmitter (ST), and a secondary receiver (SR) operating in FD mode, whereby ST harvests energy from both PT and SR during the first communication time slot. In the second time slot, SR receives its signal from ST while PT sends its signal to PR. The BEP expressions for the primary/secondary users (PU/SU) are obtained analytically and verified through Monte-Carlo simulations using both linear and NL EH models at ST. Additionally, to determine the trade-off between EH and IP, corresponding system performances are evaluated with regard to a power allocation coefficient at SR. The results demonstrate that, in contrast to the non-cooperative (direct transmission) case, the proposed FD-CR system with NL EH improves the PU BEP performance. Besides, SU benefits from the licensed spectrum of PU with significant BER performance.

In the second part of the thesis, the performance of a wireless-powered, two-hop, amplify-and-forward relaying system is studied when there is no direct link between the source and the destination. The power-constrained source and relay receive energy from a dedicated power beacon (PB) that broadcasts an energy-bearing signal. For both linear and NL energy harvesting models, theoretical derivations of BEP, outage probability, and throughput expressions are performed. Additionally, Monte Carlo simulations are performed to verify the theoretical results that are presented for various system parameters. The results show how the realistic NL EH model is different from the traditional linear EH model, which overestimates the performance of the system when a large amount of energy is harvested. This results in a misunderstanding of the actual performance of EH systems. However, both models operate similarly and provide appropriate results at low levels of harvested energy.

In the last part of this dissertation, the performance of the proposed two novel NL EH models is analyzed in terms of average harvested power, throughput, and BEP. The system comprises a single multi-antenna power-constraint source that transmits its signal to a destination with multiple antennas while harvesting power from a dedicated PB. For a comprehensive analysis of the system, closed-form expressions are derived for Nakagami- m fading channels and the special case of Rayleigh channels. In addition, for existing NL EH models, the simulation results are obtained using the Monte-Carlo method. The results provide a broader picture of EH systems and comprehensively compare the proposed NL EH models to linear, piece-wise linear, and NL EH models available in the literature. As a result, these provide better perspectives on the analysis and design of EH systems.

YENİ NESİL TELSİZ İLETİŞİM SİSTEMLERİ İÇİN YENİ RF ENERJİ HASATLAMA MODELLERİ

ÖZET

Yeni nesil telsiz iletişimde, birden fazla antene sahip çok sayıda düğümün birbirine bağlanması beklenmektedir. Genelde bu düğümler, kapasitelerini ve çalışma sürelerini sınırlayan bir pil ile çalışır. Fakat, bu durum enerji sınırlamalarına neden olması gibi çeşitli zorluklar oluşturmaktadır. Başka bir deyişle, piller enerji sınırlı olduğundan, bu düğümlerin enerji kısıtlaması bir zorluk oluşturmaktadır. Enerji hasatlama (EH), telsiz iletişim sisteminin ömrünü uzatabilecek umut vadeden bir yaklaşım olarak kabul edilmektedir. Bu teknoloji aynı zamanda, bu düğümlerde bulunan pillerin değiştirilmesini gerektirmediğinden yeşil iletişim endişelerini de giderir. Dolayısıyla bu düğümlerin çalışma süresini uzatır. EH’de rüzgar, güneş, titreşim vb. kaynaklar kullanılarak enerji elde edilir. Böylece hasatlanan enerji elektriğe dönüştürülür ve istenilen düğümler tarafından kullanılabilir. Ancak, yukarıda bahsedilen geleneksel EH yöntemleri her zaman mevcut değildir.

Radyo frekansı (RF) EH yöntemleri, bu düğümlerin enerji sorununu çözmek için gelecek vaat eden çözümler arasında sayılmaktadır. Kısıtlı enerji kaynakları ve çevreci iletişim göz önüne alındığında, RF EH, sensör düğümlerinde pil ve elektronik güç kaynaklarının kullanımının yerini aldığından enerji açığını çözmek için bir alternatiftir. EH düğümü, gelen işaret enerjisini hasatlama yeteneği sağlayan bir EH devresi ile donatılmıştır. Telsiz güç aktarımı (WPT) ve eşzamanlı telsiz bilgi ve güç aktarımı (SWIPT), pilsiz düğümlerin tasarımında dikkate alınan iki EH yöntemidir. EH devre özelliğine bağlı olarak güç kısıtlı düğüm gelen toplam işaret gücünü hasatlayabilir veya aynı anda enerjiyi hasatlayıp bilgiyi işleyebilir. Güç bölünmesi (PS) ve zamanda anahtarlama (TS), SWIPT yönteminin örnekleridir. Güç kısıtlı düğüm, işaret gücünün bir kısmını enerji toplamak için ve geri kalanını bilgi işleme için kullanmak üzere PS EH modunda gelen işaret enerjisinden güç toplar. Diğer yandan, TS EH modunda, EH ve bilgi işlemenin (IP) her birine, örtüşmeyen iki zaman aralığı atanır. WPT’de özel güç işareti (PB) veya çevresel güç düğümleri, güç kısıtlı düğümlerin güçlerini topladığı kaynaklardır.

Yukarıda bahsedilen EH devreleri ve EH kaynaklarında EH sistem performansını doğrudan etkileyen faktörler arasında hasatlanan gücün miktarı ve EH devresinin özelliği bulunmaktadır. Doğrusal, parçalı-doğrusal ve doğrusal olmayan EH (NL-EH) modeller, hasat edilen gücün miktarını belirlemektedir. Literatürde yer alan pek çok çalışma, EH devresinin giriş ve çıkış güçleri arasında doğrudan bir orantıya sahip olan doğrusal EH modelini dikkate almaktadır. Başka bir deyişle, doğrusal EH modeli için, hasat edilen güç doğrusal olarak EH devresinin giriş gücüne sabit bir

oran parametresiyle, yani dönüşüm verimliliğiyle bağlıdır. Öte yandan, düşük giriş gücü bölgesi için doğrusal model ile aynı miktarda enerji sağlayan parçalı doğrusal EH modeli literatürde ele alınmaktadır. Parçalı doğrusal EH modelinde, doğrusal EH modelinden farklı olarak, yüksek giriş güçlerinde çıkış gücü sabit bir maksimum değerle sınırlanmaktadır. Bununla birlikte, tasarlanan pratik EH devrelerinden alınan veri ölçümlerine dayanarak, doğrusal EH modelleri, özellikle yüksek giriş güçleri için hasat edilen enerji miktarını olduğundan fazla tahmin eder ve bu da sistem performansının yanlış anlaşılmasına neden olur. Ayrıca, parçalı-doğrusal EH yöntemi de genel bir bakış açısı sağlamaktadır ve kesin sistem performansı elde edilemez. Literatürde, pratik deneysel testlerden elde edilen verilerle, EH devresinin giriş ve çıkış güçleri ilişkisinin doğrusal olmadığı ve giriş gücünü artırsa bile çıkış gücünün maksimum güce sabitlendiği anlaşılmaktadır. Literatürde ölçüm verileri üzerine eğri uydurma uygulanması bazı NL-EH modellerinde önerilmektedir. Bilinen bu NL-EH modelleri göz önüne alındığında, sistem performansının teorik ifadelerinin matematiksel olarak işlenebilir olmadığı unutulmamalıdır. Bununla birlikte, doğrusal ve parçalı doğrusal EH modelleri matematiksel olarak işlenebilir ve sistemin genel ve basit bir analizini sağlar.

Bu tez, IoT ve sensör düğümleri gibi düşük enerji iletim düğümlerinde RF EH ile ilgili pratik bir konsept geliştirmektedir. Bu tezin amaçları aşağıdaki gibi sıralanabilir: Yeşil iletişimi dikkate alan yeni nesil telsiz iletişim sistemlerinin gereksinimlerini karşılamak için RF EH alanında getirilebilecek yenilikleri araştırmak, teorik ifadelerin ve pratik gerçekçi senaryoların türetilmesine katkıda bulunmak için farklı RF EH modellerini araştırmak, sistem performansını bit hata olasılığı (BEP) ve kesinti olasılığı açısından değerlendirmek, IoT ve sensörler gibi düşük veri seviyeli iletim düğümlerinde kullanılmayı amaçlayan RF EH'yi mevcut RF EH modelleri açısından tartışmak ve ardından matematiksel olarak işlenebilir NL EH modelleri önermektir. Yukarıda açıklanan bağlamda, bu tezin ilk adımı, literatürde mevcut EH modelleri ile farklı sistemlerin başarımlarını araştırmaktır. İkinci olarak, bu tezde, EH sistem tasarımına geniş bir bakış açısı sağlayan matematiksel olarak işlenebilir EH modelleri önerilmektedir.

Tezde, üç farklı sistem modeli incelenmiş ve bu RF EH sistemlerinin performansları için kapalı formda analitik ifadeler türetilmiştir. Ayrıca, göz önüne alınan her sistemin performansı, BEP ve kesinti olasılığı açısından değerlendirilmiştir. Bunun dışında, farklı sistem parametreleri için teorik sonuçlar elde edilmiş ve Monte-Carlo simülasyon yöntemi ile doğrulanmıştır. Bu nedenle, bu tez kapsamında yapılan çalışmaların sonuçlarının, EH sistem tasarımına bakış açısını geliştirecek olması ve üzerinde çalışılan sistemlere ilişkin kapsamlı bilgiler sağlayacak olması açısından günümüz literatürüne önemli katkılarda bulunacağı öngörülmektedir.

Tezin ilk bölümünde, doğrusal/doğrusal olmayan EH yeteneğine sahip tam çift yönlü (FD) üzerine sermeli (overlay) bilişsel radyo (CR) ağının bit hata oranı (BER) performansı incelenmiştir. Ele alınan üzerine sermeli CR ağı, bir birincil verici/alıcı (PT/PR) çifti, bir ikincil verici (ST) ve ST'nin ilk iletişim süresi boyunca hem PT hem de SR'den enerji hasatladığı FD modunda çalışan ikincil alıcıdan (SR) oluşmaktadır. İkinci iletişim zaman diliminde SR, işaretini PR'ye ileterek PT ile işbirliği yaparken ST'den gelen işareti almaktadır. Birincil/ikincil kullanıcılar (PU/SU) için bit hata olasılığı (BEP) ifadeleri, ST'de uygulanan hem doğrusal hem de doğrusal olmayan EH modellerine göre analitik olarak türetilmekte, Monte-Carlo simülasyon sonuçlarıyla

doğrulanmaktadır. Ayrıca hem PU hem de SU'nun BER'ini en aza indirme hedefi göz önünde bulundurularak PT ve ST arasındaki optimum mesafe değeri belirlenmektedir. Ayrıca, EH ve bilgi işleme arasındaki dengeyi elde etmek için SR'de bir güç paylaşım katsayısına bağlı olarak sistem performansları elde edilmektedir. Sonuçlar, doğrusal olmayan EH'ye sahip önerilen FD-CR sisteminin, işbirlikli olmayan (doğrudan iletim) duruma kıyasla PU BEP performansını iyileştirdiğini, SU'nun ise yeterli bir BER performansı ile PU spektrumundan yararlandığını göstermektedir.

Tezin ikinci çalışması olarak, kaynak ve hedef arasında doğrudan bir bağlantının olmadığı durumda, telsiz iki atlamalı (DH) yükselt ve ilet (AF) tekniği ile iletim yapılan bir sistem ele alınmaktadır. PB nin, gücü kısıtlı kaynak ve rölenin enerji hasatladığı enerji taşıyan bir işaret ilettiği varsayılmaktadır. BEP, kesinti olasılığı ve verim (throughput) ifadelerinin teorik derivasyonları, hem doğrusal hem de doğrusal olmayan enerji hasatlama modelleri için gerçekleştirilmektedir. Ayrıca, farklı sistem parametreleri için sağlanan teorik sonuçlar Monte-Carlo simülasyonları ile doğrulanmaktadır. Elde edilen sonuçlar, gerçekçi doğrusal olmayan EH modeli ile sistem performansını hasatlanan yüksek enerji seviyelerinde hatalı temsil eden geleneksel doğrusal EH modeli arasındaki farkı ortaya koymaktadır. Bununla birlikte, hasatlanan düşük enerji seviyelerinde, her iki model de benzer şekilde davranmakta ve gerçekçi sonuçlar vermektedir.

Tezin üçüncü çalışmasında, sistem performansının ortalama hasatlanan güç, sistem verimi ve bit hata olasılığı açısından incelendiği, matematiksel olarak işlenebilir iki yeni doğrusal olmayan EH (NL-EH) modeli önerilmiştir. Sistem, gücünü özel bir güç kaynağından hasatlayan güç kısıtlı çok antenli bir kaynaktan ve çok antenle donatılmış bir hedeften oluşmaktadır. Sistemin kapsamlı bir analizi için, Nakagami- m ve Rayleigh sönümlemeli kanallarının özel durumu için kapalı formda ifadeler türetilmiştir. Karşılaştırma amacıyla, literatürde verilen doğrusal ve parçalı doğrusal EH modelleri için Nakagami- m sönümlemeli kanallardaki performans ifadeleri türetilmiştir. Karşılaştırma amacıyla, literatürde var olan NL-EH modelleri için Monte-Carlo yöntemi uygulanarak simülasyon sonuçları elde edilmiştir. Sonuçlar, EH sistemlerine daha geniş bir bakış sağlamakta ve önerilen yeni modeller literatürde mevcut olan doğrusal, parçalı doğrusal ve NL EH modelleriyle karşılaştırılmaktadır.



1. INTRODUCTION

Power requirement will be a key constraint for sustainability in next-generation wireless communications, where a significant number of nodes with many antennas will be connected to one another. Moreover, a large number of batteries should be continuously replaced manually to guarantee continuous wireless communication between these nodes. This introduces another challenge which lies in the concerns of green communication. One of the potential solutions to the energy constraint challenge of these nodes is energy harvesting (EH) technology. This technology also addresses the concerns of green communication in which no batteries are needed to be replaced in those nodes. Hence, it extends the power-constraint node operation time. In this technology, the battery-free nodes harvest their power from several sources such as solar, wind, vibration, radio frequency (RF), and etc. The harvested energy is then transformed into electricity and used as a transmit power to broadcast the desired signal to the destination. However, some of the aforementioned EH methods are impractical in some harsh environments or indoor communications. Consequently, it decreases the possibility of continuous transmission. Moreover, RF EH is usually available and the power-constraint nodes harvest their power from the incoming signal from the environment. Thus, RF EH offers an alternative step to overcome the energy scarcity since it eliminates the necessity of batteries and electrical power supply at the sensor nodes.

In the literature, RF EH sources are classified into two forms: dedicated and ambient. In dedicated RF sources, a power beacon, which is especially assigned to a power-constraint node, broadcasts an energy-bearing signal. Furthermore, TV and radio transmitters, and mobile base stations are examples of the latter RF energy source. Then, the available EH circuit in power-constraint nodes allows us to capture the energy of incoming signals from these sources. Also, wireless power transfer (WPT) and simultaneous wireless information and power transmission (SWIPT) are the two EH techniques taken into consideration in the circuit design of these nodes.

The power-constraint node may either harvest all of the incoming signal power based on the WPT design of EH circuit or concurrently harvest energy and process data depending on the SWIPT EH circuit characteristic. Moreover, SWIPT methods include power-splitting (PS) and time-switching (TS) EH modes. In PS mode, the incoming signal energy is divided into two parts, one portion of the signal is used as information processing (IP), while the remainder is harvested in the circuit and used as transmitter power. In TS mode, all of the incoming energy is harvested at a specific time interval while a dedicated time interval is only used for IP. Moreover, the power-constraint nodes of the WPT acquire their energy from ambient power nodes or dedicated power beacons (PB), which transmit only the energy-bearing signals. The amount of harvested power and the characteristics of the EH circuit, which directly impact the EH system performance, are essential elements in the above-mentioned kinds of EH systems.

Different linear and non-linear (NL) EH models are presented in the literature for the performance analysis of EH systems. Models that characterize the amounts of harvested power include linear, piece-wise linear, and NL EH models. According to the linear EH model, the conversion efficiency acts as a constant ratio parameter, causing the harvested power to depend linearly on the input power of the EH circuit. However, based on data measurements from developed realistic EH circuits, the linear EH model exaggerates the amount of harvested energy, particularly for large input powers, which ultimately leads to a misconception of the system performance. Therefore, the piece-wise linear EH model is described in the literature; it behaves identically to the linear EH model for small input powers. However, at high input powers, it saturates at a predefined threshold power. Furthermore, the NL characteristics of the EH circuits may be described by a NL function of the amount of harvested power relative to the input power. With the measured data provided from practical experimental tests, it is found that the EH circuit's relationship between input and output power is NL and that the output power saturates to a maximum power even when the input power is increased. Some NL-EH models are presented in the literature using curve fitting over measurement data. This feature of the EH node differs from the linear EH model, in which the harvested power is directly proportional to the input power and increases as the input power of the EH circuit increases. As a consequence, the amount

of harvested power is overestimated, which leads to a misinterpretation of the EH system's performance.

Even with these well-known NL-EH models in the literature, the theoretical derivations of system performance are not mathematically comprehensible. The linear and piece-wise linear EH models, however, provide a general and straightforward analysis of the EH system and are mathematically tractable. However, the NL-EH models that are proposed in this thesis enable the theoretical derivation of system performance since they are mathematically tractable. This gives an advantage in the design and evaluation of EH systems. Hence, this provides engineers with comprehensive insight through the EH system capability of the Internet of Things (IoT) nodes.

In the context of this thesis, different system models considering various EH modes and models are analyzed, and their performances are evaluated. Moreover, two practical NL-EH models are proposed and compared with the available EH models in the literature. Thus, this provides a comprehensive analysis of the EH system design.

1.1 Purpose of Thesis

The requirement for external power supply increases significantly as the demand for IoT and sensor nodes rises. Batteries are finite resources with a harmful impact on the environment, in addition to the issues as charging, replacing, size, and weight. Therefore, finding an alternative approach to overcome these power restrictions is highly desirable. Thus, using the available RF waves in the environment, power-constraint nodes can be charged automatically, which provides continuous data transmission between sensors. This thesis develops a practical concept regarding RF EH in low energy transmission nodes such as the IoT and sensor nodes. The aim of this thesis is as follows:

- Investigate several studies in the field of RF EH to meet the requirements of next generation wireless communication systems considering green communication.
- Investigate different RF EH models to contribute to the derivation of theoretical expressions and practical realistic scenarios.

- Evaluate the system performance in terms of bit error probability (BEP) and outage probability.
- Discuss the RF EH which aims to be used in low data level transmission nodes like IoT and sensors, from the perspective of available RF EH models and then propose mathematically tractable NL EH models.

In the context described above, the first step of this dissertation is to investigate the system performance of different proposed system models with available EH models in the literature. Secondly, in this thesis, we propose mathematically tractable EH models that provide a broad insight into the EH system design.

In the first part of the thesis, we combine the full-duplex (FD) EH system's capabilities with overlay cognitive radio (CR) and analyze the bit error rate (BER) performance of the proposed system for both linear and NL EH models. We apply the piece-wise linear EH model to analyze the BEP performance of the FD-CR network since it simplifies the derivations of the probability density function (PDF) and cumulative distribution function (CDF) of the received signal-to-noise ratio (SNR)s. In contrast to the majority of current studies in the literature, the secondary transmitter (ST) is regarded as a power constraint node. Additionally, the FD protocol is taken into account at the secondary receiver (SR), which transmits signals from the primary user (PU) to the primary receiver (PR), an energy-bearing signal to the ST, and data from the ST. Moreover, a power allocation coefficient is taken into account, allowing SR to divide its total power into two time slots so that EH may operate with PU. The interference link between ST and PR is also taken into consideration to make the communication environment more realistic. On the other hand, the amounts of energy harvested for both linear and NL EH models are determined with respect to the input power of the PU.

In the second part of the thesis, we investigate theoretical expressions for the BEP and throughput of a wireless-powered dual-hop amplify-and-forward (WP DH AF) relaying system. Furthermore, we derive a theoretically tractable linear piece-wise EH model in terms of the PDF, CDF of harvested power, and system performance. Besides, for the linear EH model given in the literature, theoretical expressions for the performance of the proposed system are provided. Unlike the linear EH model, which misinterprets system performance, the considered linear piece-wise EH model,

along with the practical NL EH models, delivers actual system performance. Since the linear EH model has been considered in the majority of papers in the literature, this section also analyzes the linear piece-wise EH model's effects to give a thorough system analysis.

In the last part of the thesis, we provide two NL-EH models that are mathematically tractable in terms of the PDF, and CDF of harvested power, and system performance, despite a slight discrepancy between the measured data points and those predicted by our models. Unlike the NL-EH models in the literature, our models are not based on the polynomial, exponential, or cubic functions of the input power. Instead, they are based directly on the input power. The considered system consists of a PB from which a power-constrained source harvests energy, which is then used to transmit data to a destination. The PDF of harvested power is applied to calculate the average harvested power. Additionally, for the two proposed NL-EH models, the expressions for the outage probability, throughput, and BEP of the proposed system are derived in closed-forms. It should be noted that since the available NL-EH models in the literature are not mathematically tractable, they do not provide closed-form expressions for the system performance. Only the amount of power harvested at the EH nodes is analyzed in the papers that take these models into consideration; system BEP or throughput performance are not investigated.

The numerical results provide a comprehensive study of all the considered EH models in terms of system performance. The results show that the distance and number of antennas have a significant impact on the system performance, which provides insight into its maximization.

1.2 Hypothesis

The ability to harvest and store energy from radio signals has made EH an attractive topic of study in recent years, as it presents a feasible option for energy-constrained wireless communication systems. Transmission of both information and power can be achieved via radiated RF signals. As a result, EH circuits can deliver a substantial amount of energy to wireless networks. Most studies focus on a linear model of EH, where the amount of harvested energy is proportional to the amount of input power. However, the RF-to-DC rectifier of the EH circuit has constraints that make the

harvested amount of energy a NL function of the input power.. This dissertation can be considered as a theoretical investigation of EH behavior in battery-free nodes. As part of the theoretical analysis, the proposed systems are investigated with the available RF EH models provided in the literature. Moreover, two practical NL-EH models are proposed and compared with the conventional EH models. The theoretical derivations are obtained as closed-form expressions in the next chapters.

The contributions of this thesis are listed as follows:

1. In the first part of the thesis, an overlay CR system with FD capability is considered and the contributions are listed as follows:
 - The theoretical derivations for the BEP of PU and SU are obtained at ST applying the NL and linear EH models.
 - The effects of distance on the performance of both the PU and SU systems are analyzed.
 - The system performances of NL EH model are investigated by applying a predefined threshold power and compared with those of the linear EH model and direct transmission.
2. In the second part of the thesis, a DH EH system model where both source and relay are power-constraint is investigated and the contributions are listed as follows:
 - Closed-form BEP expressions are theoretically derived using both linear and NL EH models.
 - The throughput of considered system is calculated under both linear and NL EH models.
 - Various system parameters, including distance, power, and achievable rate, are considered along with simulation and theoretical results for the considered system.
3. In the third part of the thesis, based on the curve-fitting of experimental data two mathematically tractable NL EH models are proposed and compared with available EH models in the literature. The contributions of the considered system are listed as follows:

- For the proposed NL-EH models, PDF and CDF of the harvested power are provided.
- The average harvested power, throughput, and BEP are theoretically derived and the numerical results are supported by Monte-Carlo simulations.
- The system performance is theoretically derived for both Nakagami- m and Rayleigh fading channels.
- Results are compared with existing linear, piece-wise linear, and NL-EH models from the literature.





2. BER PERFORMANCE OF FULL-DUPLEX COGNITIVE RADIO NETWORK WITH NONLINEAR ENERGY HARVESTING¹

In this chapter, the bit error rate (BER) performance of a full-duplex (FD) overlay cognitive radio (CR) network with linear/nonlinear energy harvesting (EH) capability is investigated. The considered overlay CR network consists of a primary transmitter/receiver (PT/PR) pair, a secondary transmitter (ST) and an FD mode operated secondary receiver (SR) where ST harvests energy from both PT and SR during the first communication time slot. In the second communication time slot, SR receives its intended signal from ST while cooperating with PT by transmitting its signal to PR. The bit error probability (BEP) expressions for the primary/secondary users (PU/SU) are analytically derived and validated through Monte-Carlo simulations according to both linear and nonlinear EH models applied at ST. In addition, the optimum distance value between PT and ST is determined considering the objective of minimizing BER of both PU and SU. Moreover, corresponding system performances are obtained regarding a power allocation coefficient at SR to obtain the trade-off between EH and information processing. The results show that the proposed FD-CR system with nonlinear EH improves the PU BEP performance compared to the non-cooperative (direct transmission) case while the SU benefits from the spectrum of PU with a significant BER performance.

2.1 Introduction

In recent years, energy harvesting (EH) has taken great attraction since it offers a promising solution for energy-constraint wireless communication systems thanks to the aggregation and accumulation of energy from radio signals. Through the radio frequency (RF) signal radiation, not only information transmission but also power transmission can be carried out [4,5]. Thus, a reasonable amount of energy for wireless

¹This chapter is based on "Babaei, M., Aygözü, Ü., Basaran, M. and Durak-Ata, L. (2020). BER Performance of Full-Duplex Cognitive Radio Network With Nonlinear Energy Harvesting, IEEE Trans. Green Commun. Netw., vol. 4, no. 2, pp. 448-460, doi: 10.1109/TGCN.2020.2990882."

networks can be provided by EH. In the literature, most of the works are concerned with a linear model of EH in which harvested amount of energy increases linearly with the input power. However, in practice due to the constraints in RF-to-DC rectifier of the EH circuit, harvested amount of energy is nonlinear [6–10].

Considering the unforeseen exponential growth in the demand of both mobile applications and reliability of wireless communication systems besides the scarcity of the available spectrum, cognitive radio (CR) has been developed to make use of the licensed primary user's (PU) spectrum by the unlicensed secondary users (SU) under certain constraints [11]. CR provides the efficient usage of the available spectrum through dynamic spectrum access strategies which are mainly known as interweave, underlay and overlay paradigms [12]. In the interweave paradigm, SU detects the spectral holes which are not in use by PU and opportunistically intervenes during the availability of the spectrum resources. In the underlay paradigm, PU and SU may simultaneously occupy the spectrum under the condition that the interference caused by the secondary transmitter (ST) at the primary receiver (PR) is below on acceptable threshold level. Finally, in the overlay paradigm, SU shares the PU's spectrum if it cooperates with PU to improve its performance [12].

Full-duplex (FD) communication systems are proposed to support both transmission and reception of communication signals simultaneously unlike commonly used half-duplex (HD) systems in which only one operating mode of antenna (i.e., transmission or reception) is active during a signaling interval [13]. FD techniques are able to increase the capacity of cooperative wireless communication systems significantly.

In this chapter, we consider the system model as a combination of FD with nonlinear EH approaches in overlay CR networks by providing the BER performance analysis where the missing concepts in the previous works and main contributions will be defined in the following subsections.

2.1.1 Related work

Prior research studies focusing on conventional HD systems with linear EH model are given in [14–22]. In [14], the time-switching (TS) and the power-splitting (PS) relaying protocols are considered to make the relay capable of EH and information processing

(IP) with amplify-and-forward (AF) relaying technique. In [15], the expressions of average bit-error-probability (BEP) and the outage probability are derived to evaluate the performance of simultaneous wireless information and power transfer (SWIPT) technique in cooperative AF relaying communication systems by applying binary phase shift keying (BPSK) and differential BPSK (DBPSK) modulations under shadowed Nakagami- m channel fading. Similarly, the average symbol error probability (SEP), the achievable diversity order, and the throughput for a wireless EH with three-node decode-and-forward (DF) relaying system operating under Nakagami- m fading are analyzed to optimize power and time for the corresponding slot duration in [16]. In [17], the BEP performance of SWIPT-based differential AF relaying in Nakagami- m fading and a line-of-sight (LoS) link is investigated, and a closed-form expression for the probability density function (PDF) of the signal-to-noise-ratio (SNR) of the two-hop link is approximated for high SNR region. In addition, a closed-form expression of average bit error rate (BER) is derived for a differential cooperative system employing AF relaying with SWIPT capability when a selection combining scheme is applied [18]. Similar to [18], the effect of differential modulation on the system performance of SWIPT cooperative AF networks is demonstrated, and the PDF of end-to-end SNR and the corresponding average BEP are derived in [19]. On the other hand, a non-coherent SWIPT technique for DF relaying systems is proposed to bypass the need for instantaneous channel state information (CSI), and hence, to save energy in the network unlike the conventional SWIPT schemes that require the instantaneous CSI for coherent information transmission [20]. PS, TS and ideal EH protocols are considered for a power-constraint relay with the links exposed to Nakagami- m fading and the corresponding results are provided for different system parameters [21]. Furthermore, BER performance of DH network with AF and DF relaying protocols is analyzed in [21]. In [22], the system performance of selective-DF (SDF) with direct link from source to destination is investigated where PS, TS and ideal EH models are applied at the relay.

Furthermore, recent studies concentrating on FD communication with EH are given in [23–30]. In [23], the outage performance of FD SWIPT with DF relaying is investigated for the case where the self-loop channel suffers from fading. In [24], FD two-way (TW) relay networks with EH at relays are considered where

the throughput and outage probability expressions for delay limited transmission in outdoor environment are derived. On the other hand, the wireless-powered FD TW cooperative system applying spatial modulation is considered in [25]. Without any limitation on the number of cooperating relays, both single relay selection and general relay selection (GRS) techniques are investigated in [26] where the authors propose several low-complexity heuristic GRS methods based on various relay ordering and greedy selection criteria. In [27], optimal TS is studied for FD dual-hop (DH) relaying system with EH to improve the achievable throughput. The effect of noise power and distance on the FD cooperative AF and DF relaying systems based on RF EH is investigated in terms of the outage probability and throughput in [28]. In addition, a novel FD two-phase protocol for SWIPT is proposed both to provide uninterrupted information transfer and to enable self-energy recycling in [29]. Different from [24], the ergodic outage probability performance of DH, HD and FD, AF and DF relaying with EH is analyzed over indoor channels characterized by log-normal fading in [30].

Note that, all the previous works mentioned above assume the linear EH model with a fixed energy conversion efficiency [14–30]. More simply, energy conversion efficiency is assumed to be independent of the input power of the EH circuit and the amount of the harvested energy is linearly proportional to the input power. However, in practice EH circuits have a nonlinear behavior and depend on the input power of the RF-DC circuit [6–10]. A practical nonlinear EH model based on the Logistic (Sigmoid) function which captures the characteristic of EH circuit is proposed in [6]. Moreover, an optimization problem is formulated to maximize the amount of harvested energy. A heuristic nonlinear EH model of the energy conversion efficiency depending on the input power is investigated in [7]. The rational nonlinear model is obtained after applying curve fitting, where the model is chosen based on the trade-off between accuracy and complexity of the result. In [8], the piece-wise linear model of EH receiver that captures the saturation characteristic of the practical EH circuit is proposed. A practical quadratic nonlinear EH model is studied in [10]. Goodness of the fit in [10] and [6] are compared with each other where the value of root-sum-of-squares error (RSSE) in [10] is found to be larger than that of [6]. Moreover, the practical model given in [6] is not a concave function of the input power which prevents the convex optimization. However, the authors in [10] propose a second order polynomial

as the EH model satisfying convexity properties and providing a good match with the practical measured data. In [9], the authors provide a simpler modification of the nonlinear EH model given on [7]. However, the proposed models in [6, 7, 9, 10] are not mathematically tractable to derive the PDF and cumulative distribution function (CDF) of the received SNRs and to average over the channel fading coefficients.

Recently, the outage capacity of AF relaying protocol for Nakagami- m fading channel is derived assuming nonlinear EH model at power-constraint relay in [8]. In [31], the performance of DH multiple-input multiple-output (MIMO) system with imperfect CSI and nonlinear EH model is investigated. Furthermore, PS EH and DF relaying protocols are applied at the relay. Moreover, a saturation value is provided for the nonlinear energy harvester at each relay.

2.1.2 Motivation and main contributions

Studies in [14–22] concentrate on conventional HD-EH systems while [23–30] focus on linear FD-EH systems mainly based on the throughput and outage probability analyses in addition to the optimization issues like time and power adjustments under different parameters including modulation type, relay scenario, and channel conditions.

In this chapter, we combine the FD-EH system capabilities with overlay CR and investigate the BER performance of the proposed system for both linear and nonlinear EH models [8]. In order to derive the BEP performance of FD-CR network, we adopt the nonlinear EH model given in [8] which makes tractable the calculation of PDF and CDF of the received SNRs. To the best of author's knowledge, BEP performance for the nonlinear EH model has not been investigated yet in the literature. Apart from most of the existing literature, ST is considered as power constraint. Also, FD protocol is assumed at the secondary receiver (SR) that transmits PU's data to PR, energy bearing signal to ST and receives ST's data. In addition, a power allocation coefficient, for SR dividing its total power into two time slots for EH to be cooperated with PU, is considered. Furthermore, interference link from ST to PR is taken into account that makes the communication environment more realistic. Moreover, the amounts of harvested energy for linear and nonlinear EH models are obtained with respect to the PU's input power. The main contributions of this chapter are as follows:

Table 2.1 : Parameters and notations.

Parameters	Description
${}_pF_q(a, b; c; \cdot)$	Hyper-geometric function [32, 9.10]
$K_\nu(\cdot)$	Modified Bessel function of the second kind [32, 8.446]
$G_{pq}^{mn} \left(\cdot \middle \begin{smallmatrix} a_p \\ b_q \end{smallmatrix} \right)$	Meijer-G function [32, 9.3]
$\text{erfc}(\cdot)$	Complementary error function [33, 2.3-17]
$Q(\cdot)$	Q function [33, 2.3-10]
$\Gamma(\cdot)$	Gamma function [33, 8.31]
$\Gamma(\cdot, \cdot)$	Upper Gamma function [32, 8.35]
$\gamma(\cdot, \cdot)$	Lower Gamma function [32, 8.35]
$f(\cdot)$	Probability distribution function (PDF)
$F(\cdot)$	Cumulative distribution function (CDF)
η	Energy conversion coefficient ($0 < \eta \leq 1$)
P_{pt}	Transmit power of the PT
$P_{sr} = P_{sr1} + P_{sr2}$	Transmit power of the SR
m_{43}	Channel parameter of link SR $\rightarrow$ ST
m_{34}	Channel parameter of link ST $\rightarrow$ SR
a, b	Modulation specific constants [21]

- The theoretical closed-form expressions for the BEP of PU and SU are derived according to the nonlinear and linear EH models at ST.
- Effects of the distances on both PU and SU system performances are investigated.
- System performances are analyzed using a predefined threshold for the nonlinear EH model and are compared with the linear EH model and the direct transmission.

2.1.3 Organization and notation of this chapter

The rest of this chapter is organized as follows. The system model is explained in Section 2.2. Theoretical BEP derivations are provided in Section 2.3. Numerical results are presented in Section 2.4. Concluding remarks are given in Section 2.5. The parameters and notations used throughout this chapter are listed in Table 2.1.

2.2 System Model

The considered cooperative FD EH overlay CR system model is shown in Figure 2.1 (a) and (b), where ST is equipped with an RF-EH capable device. Primary transmitter (PT), PR and ST are equipped with a single antenna. SR has two antennas operating in FD mode. h_{ij} and d_{ij} represent the channel fading coefficients and

link distances from the i^{th} receiver node to the j^{th} transmitter node, respectively, where $ij \in \{21, 31, 41, 23, 24, 33, 34, 43\}$. Moreover, links $ST \rightarrow SR$ and $SR \rightarrow ST$ are assumed stronger compared to other ones and exposed to Nakagami- m fading with $m > 1$. The remaining links are exposed to Rayleigh fading. All links are assumed as frequency flat fading channel for which the channel gains remain unchanged during each transmission interval of T seconds but changes independently from one transmission interval to another. Node numeration and transmission between nodes considering time slots are indicated in Figure 2.1 (a) and (b). Furthermore, the total transmission time T is equally divided into two equal time slots as $T/2$ seconds. Path loss component, denoted by L_{ij} , is modeled as $L_{ij} = 1/(1 + d_{ij}^\nu)$ where ν is the path loss exponent of the channel [21, 34]. $E[|h_{i,j}|^2] = \Omega_{ij}$ represents the expected channel power gain for each link. At each node, additive white Gaussian noise (AWGN) sample is independent and identically distributed (i.i.d.) as complex Gaussian where $n \sim \mathcal{CN}(0, 2\sigma^2)$ with $\sigma^2 = N_o/2$. x_{pu} and x_{su} are the transmitted signals of PU and SU, respectively, with $E[|x_{pu}|^2] = E[|x_{su}|^2] = 1$.

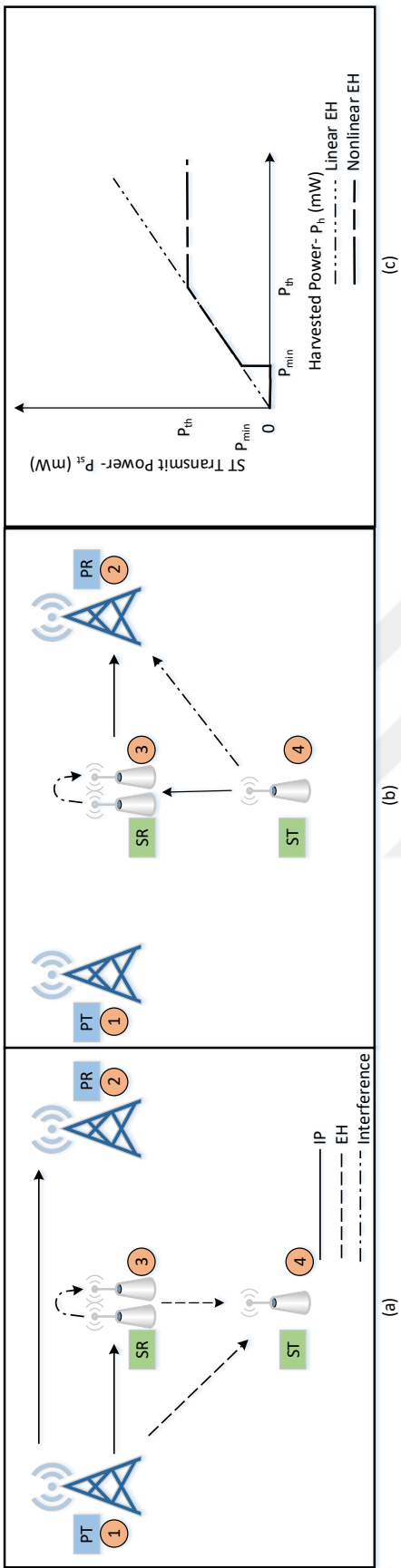


Figure 2.1 : Illustration of CR network with FD EH secondary receiver. (a) First time slot, (b) Second time slot, (c) Linear and nonlinear EH model.

2.2.1 Communication at the first time slot

At the first time slot of $T/2$ seconds illustrated in Figure 2.1 (a), PT transmits its signal to PR and at the same time SR and ST receives this signal. Moreover, during this time slot, SR transmit an energy signal to ST for EH. Since a small amount of RF energy is harvested in ST from only link PT $\rightarrow$ ST, we considered that ST harvests energy from both SR and PT which guarantees more transmit power for ST and therefore increased SU performance.

The received signal at PR can be expressed as

$$y_{pr1} = \sqrt{P_{pt}L_{21}}h_{21}x_{pu} + n_{pr1} \quad (2.1)$$

where the un-modulated energy bearing signal transmitted from SR in the first time slot can be eliminated at PR since it is converted to a direct current (DC) level after demodulation. The received SNR at PR can be given from (2.1) as

$$\gamma_{21} = P_{pt}L_{21}|h_{21}|^2/\sigma^2. \quad (2.2)$$

In (2.2), γ_{21} is exponentially distributed as $\gamma_{21} \sim \text{Exp}(\bar{\gamma}_{21})$ where $\bar{\gamma}_{21} = P_{pt}L_{21}\Omega_{21}/\sigma^2$. The received signal at SR is written as,

$$y_{sr1} = \sqrt{P_{pt}L_{31}}h_{31}x_{pu} + \sqrt{P_{sr1}L_{33}}h_{LI}x_E + n_{sr1} \quad (2.3)$$

where h_{LI} and x_E are the self-interference component of SR and the energy bearing signal, respectively. Moreover, h_{LI} results from imperfect loop interference elimination [35] and is assumed to be non-fading and a fixed value as $E[|h_{LI}|^2] = \Omega_{LI}^2$ [13]. If the self-interference stemming from un-modulated energy bearing signal is completely isolated [35], the second term in the right-hand side of the (2.3) is eliminated.

Since ST is capable to harvest energy from both PT and SR, the harvested amount of power at ST is given as

$$P_h = \eta(P_{pt}L_{41}|h_{41}|^2 + P_{sr1}L_{43}|h_{43}|^2). \quad (2.4)$$

2.2.2 Communication at the second time slot

ST transmits its signal to SR while the FD SR transmits PU signal received in the previous time slot to PR. Interference occurs at PR due to the transmission of ST. The received signal at PR during the second time slot (Figure 2.1 (b)) can be given as

$$y_{pr_2} = \sqrt{P_{sr_2} L_{23}} h_{23} x_{pu} + \sqrt{P_{st} L_{24}} h_{24} x_{su} + n_{pr_2}. \quad (2.5)$$

The received SNR at PR from the link SR→PR which can be expressed from (2.5) as

$$\gamma_{23} = \frac{P_{sr_2} L_{23} |h_{23}|^2 / \sigma^2}{1 + P_{st} L_{24} \Omega_{24} / \sigma^2}. \quad (2.6)$$

Please note that since PR does not know the instantaneous value of $|h_{24}|^2$, its expectation $E[|h_{24}|^2] = \Omega_{24}^2$ is replaced in (2.6). The received signal at SR can be defined as

$$y_{sr_2} = \sqrt{P_{st} L_{34}} h_{34} x_{su} + \sqrt{P_{sr_2} L_{33}} h_{LI} x_{pu} + n_{sr_2}. \quad (2.7)$$

Furthermore, the received SNR at SR from the link ST→SR can be expressed using (2.7) as

$$\gamma_{34} = \frac{P_{st} L_{34} |h_{34}|^2 / \sigma^2}{1 + P_{sr_2} \Omega_{LI} / \sigma^2}. \quad (2.8)$$

2.2.3 Linear and nonlinear EH model

In the literature, the amount of harvested energy when linear EH model is assumed, increases with increasing input power. However, as mentioned in [6–10], linear EH model is not practical since diodes, capacitors, etc. are located in the EH circuit. When considering the analytical derivations of the PDF and CDF of received SNRs, the models proposed in [6, 7, 9, 10] are not mathematically tractable. Then, the nonlinear EH model considered in [8] is adopted where a threshold value is defined for the maximum (saturation) power of the EH mode. Moreover, in addition to the EH model in [8], a minimum sensitivity harvested power for ST transmission power is considered (Figure 2.1(c)). For this nonlinear EH model, the power at ST node is expressed as

$$P_{st} = \begin{cases} 0, & \text{if } P_h \leq P_{min} \\ P_h, & \text{if } P_{min} < P_h \leq P_{th} \\ P_{th}, & \text{if } P_h > P_{th}, \end{cases} \quad (2.9)$$

where P_{th} and P_{min} are the threshold (saturation) and minimum transmission powers of ST, respectively.

2.3 Bit Error Probability Analysis

In this section, the theoretical SEP expressions for both PU and SU are derived and then assuming Gray mapping BEP can be derived. For both users, linear and nonlinear EH models for the ST battery are considered.

2.3.1 BEP analysis for PU

2.3.1.1 PU BEP for nonlinear EH model

The overall SEP of the PU can be defined as

$$P_{pu}^{nL} = P_e^{pt \rightarrow sr} P_e^{pt \rightarrow pr} + (1 - P_e^{pt \rightarrow sr}) P_{MRC}^{nL} \quad (2.10)$$

where $P_e^{pt \rightarrow pr}$ and $P_e^{pt \rightarrow sr}$ are the SEPs of the links PT→PR and PT→SR, respectively, and can be calculated theoretically as given in [36]. In (2.10), P_{MRC}^{nL} represents the SEP of the maximal-ratio-combining (MRC) of the links PT→PR and SR→PR and can be expressed as

$$P_{MRC}^{nL} = \int_0^{P_{th}} P(e|z) f_Z(z) dz + P(e|z = P_{th}) \int_{P_{th}}^{\infty} f_Z(z) dz - \int_0^{P_{min}} P(e|z) f_Z(z) dz. \quad (2.11)$$

In (2.11), $f_Z(z)$ is the PDF of the r.v $Z = X + Y$ and is calculated in Appendix A. For the MRC case, the sum of two received SNRs at PR from the links PT→PR→ and SR→PR is written using (2.2) and (2.6) as

$$\gamma_\varepsilon = \gamma_{21} + \gamma_{23} \quad (2.12)$$

where γ_{21} and γ_{23} are exponentially distributed with $\gamma_{21} \sim \text{Exp}(\tilde{\gamma}_{21})$ and $\gamma_{23} \sim \text{Exp}(\kappa_1)$ with $\kappa_1 = \bar{R}/(1 + z\bar{w})$ where $\bar{R} = P_{sr2} L_{23} \Omega_{23} / \sigma^2$ and $\bar{w} = L_{24} \Omega_{24} / \sigma^2$. Please note that since $P_h > P_{th}$, in (2.6), we replace $P_{st} = z$. CDF of γ_ε in (2.12) is expressed as [37, 6-43]

$$F_\varepsilon(\varepsilon) = \int_0^\varepsilon F_{\gamma_{21}}(\varepsilon - \gamma) f_{\gamma_{23}}(\gamma) d\gamma \quad (2.13)$$

where $F_{\gamma_{21}}(\varepsilon - \gamma) = 1 - \exp(-(\varepsilon - \gamma)/\tilde{\gamma}_{21})$, $f_{\gamma_{23}}(\gamma) = \exp(-\gamma/\kappa_1)/\kappa_1$. Substituting in (2.13) after some mathematical simplification and using [32, 2.311], (2.13) can be expressed as

$$F_\varepsilon(\varepsilon) = 1 - (1 + 1/\kappa_1 \kappa_2) \exp(-\varepsilon/\kappa_1) + \exp(-\varepsilon/\tilde{\gamma}_{21})/\kappa_1 \kappa_2 \quad (2.14)$$

where $\kappa_2 = 1/\bar{\gamma}_{21} - 1/\kappa_1$. Using (2.14), the conditional BEP in (2.11) can be obtained as

$$P(e|z) = \frac{a\sqrt{b}}{2\sqrt{\pi}} \int_0^\infty \frac{e^{-b\varepsilon}}{\sqrt{\varepsilon}} F_\varepsilon(\varepsilon) d\varepsilon = \frac{a\sqrt{b}}{2\sqrt{\pi}} (N_1 - N_2(z) + N_3(z)) \quad (2.15)$$

where a and b are two phase/amplitude modulation specific constants given in Table 2.1 and

$$N_1 = \int_0^\infty \frac{e^{-b\varepsilon}}{\sqrt{\varepsilon}} d\varepsilon = \Gamma(0.5)/\sqrt{b}, \quad (2.16)$$

$$N_2(z) = (1 + 1/\kappa_1 \kappa_2) \int_0^\infty \frac{1}{\sqrt{\varepsilon}} \exp(-\varepsilon(b + 1/\kappa_1)) d\varepsilon,$$

$$N_3(z) = (1/\kappa_1 \kappa_2) \int_0^\infty \frac{1}{\sqrt{\varepsilon}} \exp(-\varepsilon(b + 1/\bar{\gamma}_{21})) d\varepsilon,$$

and N_1 is calculated using [32, 3.381-4]. Using [32, 3.361-2] for calculation of both $N_2(z)$ and $N_3(z)$ and substituting κ_1 and κ_2 , we have

$$N_2(z) = \frac{\Gamma(0.5)\bar{R}^{1.5}}{(\sqrt{b\bar{R}} + 1 + z\bar{w})(\bar{R} - \bar{\gamma}_{21} - \bar{\gamma}_{21}z\bar{w})}, \quad (2.17)$$

$$N_3(z) = \frac{\Gamma(0.5)\bar{\gamma}_{21}(1 + \bar{w}z)}{(\sqrt{b + 1/\bar{\gamma}_{21}})(\bar{R} - \bar{\gamma}_{21} - \bar{\gamma}_{21}z\bar{w})}. \quad (2.18)$$

In (2.11), $P(e|z = P_{th})$ can be calculated by substituting $z = P_{th}$ in (2.15). The first and third integrals at the right-hand side of (2.11) are expressed as

$$\int_0^\varsigma P(e|z) f_Z(z) dz = \frac{a\sqrt{b}}{2\sqrt{\pi}} (C_1 - C_2 + C_3) \quad (2.19)$$

where $\varsigma \in \{P_{th}, P_{min}\}$ and

$$C_1 = N_1 \int_0^\varsigma f_Z(z) dz,$$

$$C_2 = \int_0^\varsigma N_2(z) f_Z(z) dz,$$

$$C_3 = \int_0^\varsigma N_3(z) f_Z(z) dz.$$

Here, C_1 is calculated as derived in Appendix E, and C_2 and C_3 are obtained numerically. Substituting (2.19) in (2.11) for both $\varsigma = P_{th}$ and $\varsigma = P_{min}$ and using (E.2) in Appendix E for $\int_{P_{th}}^\infty f_Z(z) dz$, (2.11) can be calculated for the MRC link. Finally, (2.11) is substituted in (2.10) and the overall BEP of PU for the nonlinear EH model assuming Gray mapping and high SNR can be expressed as $P_{b,pu}^{nL} \simeq P_{pu}^{nL}/k$, where $k = \log_2(M)$, M being the modulation order.

2.3.1.2 PU BEP for linear EH model

The overall SEP of PU for linear EH model is expressed as

$$P_{pu}^L = P_e^{pt \rightarrow sr} P_e^{pt \rightarrow pr} + (1 - P_e^{pt \rightarrow sr}) P_{MRC}^L \quad (2.20)$$

where

$$P_{MRC}^L = \frac{a\sqrt{b}}{2\sqrt{\pi}} \int_0^\infty \frac{e^{-b\varepsilon}}{\sqrt{\varepsilon}} F_\varepsilon(\varepsilon) d\varepsilon = \frac{a\sqrt{b}}{2\sqrt{\pi}} (I_1 - I_2) \quad (2.21)$$

and $F_\varepsilon(\varepsilon)$ is the CDF of γ_ε in (2.12) which is defined in (2.13). Please note that for the linear EH model, γ_{23} is different from the nonlinear EH model. After some mathematical simplification, CDF can be rewritten as

$$F_\varepsilon(\varepsilon) = F_{\gamma_{23}}(\varepsilon) - \exp(-\varepsilon/\bar{\gamma}_{21}) \int_0^\varepsilon \exp(\gamma/\bar{\gamma}_{21}) f_{\gamma_{23}}(\gamma) d\gamma. \quad (2.22)$$

Here, $F_{\gamma_{23}}(\gamma)$ and $f_{\gamma_{23}}(\gamma)$ are calculated in Appendix B and Appendix C, respectively. Moreover, by substituting $F_\varepsilon(\varepsilon)$ from (2.22) in (2.21), assuming that

$$I_1 = \int_0^\infty \frac{e^{-b\varepsilon}}{\sqrt{\varepsilon}} F_{\gamma_{23}}(\varepsilon) d\varepsilon \quad (2.23)$$

and

$$I_2 = \int_0^\infty \frac{e^{-\rho\varepsilon}}{\sqrt{\varepsilon}} T(\varepsilon) d\varepsilon \quad (2.24)$$

where $I_1 = I_{11} - I_{12} + I_{13}$, $T(\varepsilon) = \int_0^\varepsilon \exp(\gamma/\bar{\gamma}_{21}) f_{\gamma_{23}}(\gamma) d\gamma$, $\rho = b + 1/\bar{\gamma}_{21}$ and $T(\varepsilon)$ is calculated in Appendix D. Here,

$$I_{11} = \int_0^\infty \frac{e^{-b\gamma}}{\sqrt{\gamma}} d\gamma = \sqrt{\pi/b}, \quad (2.25)$$

$$I_{12} = A_1 \int_0^\infty \frac{e^{-b\gamma}}{\sqrt{\gamma}} e^{-\gamma/\bar{R}} \left(\frac{\gamma}{\bar{R}} + \nabla\right)^{-1} d\gamma, \quad (2.26)$$

$$I_{13} = A_2(j)\Gamma(j+1) \int_0^\infty \frac{e^{-b\gamma}}{\sqrt{\gamma}} e^{-\gamma/\bar{R}} \left(\frac{\gamma}{\bar{R}} + \xi\right)^{-(j+1)} d\gamma \quad (2.27)$$

where $\nabla = \psi_{41}/\bar{w}$, $\xi = \psi_{43}/\bar{w}$. (2.25) is calculated using [32, 3.381-4]. After some mathematical simplification, (2.26) and (2.27) are calculated using [32, 3.383-10] and [38], respectively as

$$I_{12} = A_1 \bar{R} \Gamma(0.5) \Gamma(0.5, \varkappa \lambda) \exp(\varkappa \lambda) / \sqrt{\varkappa} \quad (2.28)$$

where $\varkappa = \bar{R}\psi_{41}/\bar{w}$, $\lambda = b + 1/\bar{R}$ and

$$\begin{aligned} I_{13} = & A_2(j)\bar{R}^{j+1} \left[\Gamma(0.5)\Gamma(j+0.5)\zeta^{-j-0.5} \right. \\ & \times {}_1F_1(0.5, 0.5-j, \lambda\zeta) + \Gamma(j+1) \\ & \left. \times \Gamma(-j-0.5)\lambda^{j+0.5} {}_1F_1(j+1, j+1.5, \lambda\zeta) \right] \end{aligned} \quad (2.29)$$

where $\zeta = \bar{R}\psi_{43}/\bar{w}$. Using (2.25), (2.28) and (2.29), I_1 can be obtained. $I_2 = I_{21} + I_{22}$ can be expressed over the summation of its components given below as

$$I_{21} = \int_0^\infty \frac{e^{-\rho\varepsilon}}{\sqrt{\varepsilon}} G_1(\varepsilon) d\varepsilon, \quad (2.30)$$

$$I_{22} = \int_0^\infty \frac{e^{-\rho\varepsilon}}{\sqrt{\varepsilon}} G_2(\varepsilon) d\varepsilon. \quad (2.31)$$

Substituting $G_1(\varepsilon)$ and $G_2(\varepsilon)$ from (D.6), (D.7) and using [32, 7.813-1], (2.30) and (2.31) are obtained as

$$\begin{aligned} I_{21} = & \frac{1}{\bar{R}} \sum_{k=0}^\infty \frac{\phi^k}{k!} \rho^{-k-1.5} \left[\frac{A_1}{\nabla} G_{32}^{13} \left(\frac{1}{\rho\nabla\bar{R}} \middle| \begin{matrix} -k-0.5, 0, -k \\ 0, -k-1 \end{matrix} \right) \right. \\ & \left. - \frac{A_2(j)}{\xi^{j+1}} G_{32}^{13} \left(\frac{1}{\rho\xi\bar{R}} \middle| \begin{matrix} -k-0.5, -j, -k \\ 0, -k-1 \end{matrix} \right) \right] \end{aligned} \quad (2.32)$$

and

$$\begin{aligned} I_{22} = & \frac{1}{\bar{R}} \sum_{k=0}^\infty \frac{\phi^k}{k!} \rho^{-k-1.5} \left[\frac{A_1}{\nabla^2} G_{32}^{13} \left(\frac{1}{\rho\nabla\bar{R}} \middle| \begin{matrix} -k-0.5, -1, -k \\ 0, -k-1 \end{matrix} \right) \right. \\ & \left. - \frac{A_2(j)}{\xi^{j+2}} G_{32}^{13} \left(\frac{1}{\rho\xi\bar{R}} \middle| \begin{matrix} -k-0.5, -j-1, -k \\ 0, -k-1 \end{matrix} \right) \right]. \end{aligned} \quad (2.33)$$

Substituting (2.32) and (2.33), I_2 can be calculated. Furthermore, substituting I_1 and I_2 in (2.21), a closed-form expression for the SEP of MRC link can be obtained. Finally, (2.20) is calculated by using (2.21).

2.3.2 BEP analysis for SU

2.3.2.1 SU BEP for nonlinear EH model

The overall SEP of SU for the link ST $\rightarrow$ SR when ST has a nonlinear EH model is expressed as

$$\begin{aligned} P_{su}^{nL} = & \int_0^{P_{th}} P(e|z) f_Z(z) dz + P(e|z = P_{th}) \int_{P_{th}}^\infty f_Z(z) dz \\ & - \int_0^{P_{min}} P(e|z) f_Z(z) dz. \end{aligned} \quad (2.34)$$

For SU, the SNR at SR given in (2.8) is in the form $\gamma_{34} = ZU$ which the pdf of Z is given in Appendix A and U is gamma distributed whose CDF is given as

$$F_U(u) = \psi_{34}^{m_{34}} u^{m_{34}-1} \exp(-\psi_{34}u) / \Gamma(m_{34}) \quad (2.35)$$

where $\psi_{34} = m_{34}/\bar{U}$, $\bar{U} = L_{34}\Omega_{34}/(\sigma^2 + P_{sr2}\Omega_{LI})$. For the direct transmission, the conditional SEP is given as $P_s(e|u, z) = aQ(\sqrt{2b\gamma_{34}})$ [36]. We use [39] to express $Q(\cdot)$ as Meijer-G function. Finally, after applying [32, 9.31-1] we have

$$P_s(e|u, z) = (a/2\sqrt{\pi})G_{23}^{21} \left(buz \left| \begin{matrix} 0, 1 \\ 0, 0.5, 0 \end{matrix} \right. \right). \quad (2.36)$$

Averaging over the pdf of U , the SEP conditioned on Z is obtained as

$$P_s(e|z) = \int_0^\infty P(e|u, z) f_U(u) du. \quad (2.37)$$

Using [32, 7.813-1] we have

$$P_s(e|z) = aG_{33}^{22} \left(bz/\psi_{34} \left| \begin{matrix} 1 - m_{34}, 0, 1 \\ 0, 0.5, 0 \end{matrix} \right. \right) / (2\Gamma(m_{34})\sqrt{\pi}). \quad (2.38)$$

$P(e|z = P_{th})$ in (2.34) is calculated by replacing $z = P_{th}$ in (2.38). Furthermore, the first and third integrals at the right-hand side of (2.34) are written as

$$\int_0^\varsigma P(e|z) f_Z(z) dz = \Delta(P_{s1} - P_{s2}) \quad (2.39)$$

where $\varsigma \in \{P_{th}, P_{min}\}$, $\Delta = a\rho/(2\Gamma(m_{34})\sqrt{\pi})$ and $\rho = \alpha\Gamma(m_{43})/\beta^{m_{43}}$ is calculated in Appendix A. Substituting $P(e|z)$ and $f_Z(z)$ in (2.39) and using [32, 1.211-1], we have

$$P_{s1} = \sum_{i=0}^\infty \frac{(-\psi_{41})^i}{i!} \int_0^\varsigma z^i G_{33}^{22} \left(bz/\psi_{34} \left| \begin{matrix} 1 - m_{34}, 0, 1 \\ 0, 0.5, 0 \end{matrix} \right. \right) dz, \quad (2.40)$$

$$P_{s2} = \sum_{i=0}^\infty \frac{(-\psi_{43})^i}{i!} \sum_{j=0}^{m_{43}-1} \frac{\beta^j}{j!} \times \int_0^\varsigma z^{i+j} G_{33}^{22} \left(bz/\psi_{34} \left| \begin{matrix} 1 - m_{34}, 0, 1 \\ 0, 0.5, 0 \end{matrix} \right. \right) dz. \quad (2.41)$$

Using [40, 26] in (2.40) and (2.41), we have

$$P_{s1} = \sum_{i=0}^\infty \frac{(-\psi_{41})^i}{i!} \varsigma^{i+1} G_{44}^{23} \left(b\frac{\varsigma}{\psi_{34}} \left| \begin{matrix} -i, 1 - m_{34}, 0, 1 \\ 0, 0.5, 0, -(i+1) \end{matrix} \right. \right), \quad (2.42)$$

$$P_{s2} = \sum_{i=0}^\infty \frac{(-\psi_{43})^i}{i!} \sum_{j=0}^{m_{43}-1} \frac{\beta^j}{j!} \times \varsigma^{i+j+1} G_{44}^{23} \left(b\varsigma/\psi_{34} \left| \begin{matrix} -i-j, 1 - m_{34}, 0, 1 \\ 0, 0.5, 0, -(1+i+j) \end{matrix} \right. \right). \quad (2.43)$$

Substituting both $\varsigma = P_{th}$ and $\varsigma = P_{min}$ in (2.42) and (2.43) in (2.39) and then (2.39) in (2.34), the first and third integrals at the right-hand side of (2.34) are obtained. Moreover, $\int_{P_{th}}^{\infty} f_Z(z)dz$ in (2.34) is calculated in Appendix E. Finally, the BEP of SU for the non-linear EH model is obtained using (2.34) as $P_{b,su}^{nL} \simeq P_{su}^{nL}/k$.

2.3.2.2 SU BEP for linear EH model

The overall SEP of SU for the link ST→SR when ST has a linear EH model is expressed as

$$P_{su}^L = \int_0^{\infty} P_s(e|\gamma) f_{\gamma_{34}}(\gamma) d\gamma \quad (2.44)$$

where $P_s(e|\gamma) = aQ(\sqrt{2b\gamma})$ and $f_{\gamma_{34}}(\gamma)$ is the PDF of γ_{34} which can be rewritten from (2.8) in the form

$$\gamma_{34} = (X + Y)U = ZU \quad (2.45)$$

where X is exponentially distributed while Y and U have both gamma distribution. Therefore, the PDF of γ_{34} is defined as [37]

$$f_{\gamma_{34}}(\gamma) = \int_0^{\infty} \frac{1}{u} f_Z\left(\frac{\gamma}{u}\right) f_U(u) du \quad (2.46)$$

where the PDFs of Z and U are calculated in Appendix A as (A.3) and given in (2.35), respectively. Substituting (A.3) and (2.35) in (2.46), the PDF of received SNR at SR is simplified as

$$f_{\gamma_{34}}(\gamma) = B_1 - B_2 \quad (2.47)$$

where

$$B_1 = \Xi \int_0^{\infty} u^{m_{34}-2} e^{-\psi_{41}\gamma/u - \psi_{34}u} du, \quad (2.48)$$

$$B_2 = \Xi \sum_{j=0}^{m_{43}-1} \frac{\beta^j}{j!} \gamma^j \int_0^{\infty} u^{m_{34}-j-2} e^{-\psi_{43}\gamma/u - \psi_{34}u} du \quad (2.49)$$

with $\Xi \triangleq \rho \psi_{34}^{m_{34}} / \Gamma(m_{34})$. Applying the definition in [32, 3.471-9], B_1 and B_2 in (2.48) and (2.49), respectively, are finally expressed as

$$B_1 = 2\Xi \left(\frac{\psi_{41}\gamma}{\psi_{34}} \right)^{\Lambda} K_{m_{34}-1} (2\sqrt{\gamma\psi_{41}\psi_{34}}), \quad (2.50)$$

$$B_2 = 2\Xi \sum_{j=0}^{m_{43}-1} \frac{\beta^j}{j!} \gamma^j \left(\frac{\psi_{43}\gamma}{\psi_{34}} \right)^{\Lambda-0.5j} K_{m_{34}-j-1} (2\sqrt{\gamma\psi_{43}\psi_{34}}) \quad (2.51)$$

where $\Lambda = 0.5(m_{34} - 1)$. For further calculation, we express $K(\cdot)$ in both (2.50) and (2.51) as a Meijer-G function [40, 26]. Moreover, we apply [32, 9.31-1] and after some simplification (2.50) and (2.51) are re-expressed as

$$B_1 = \Xi \left(\frac{\psi_{41}}{\psi_{34}} \right)^\Lambda \gamma^\Lambda G_{13}^{21} \left(\gamma \psi_{41} \psi_{34} \middle| \begin{matrix} 0 \\ \Lambda, -\Lambda, 0 \end{matrix} \right) \quad (2.52)$$

and

$$B_2 = \Xi \sum_{j=0}^{m_{43}-1} \frac{\beta^j}{j!} \left(\frac{\psi_{43}}{\psi_{34}} \right)^{\Lambda-0.5j} \gamma^{\Lambda+0.5j} \times G_{13}^{21} \left(\gamma \psi_{43} \psi_{34} \middle| \begin{matrix} 0 \\ \Lambda-0.5j, -(\Lambda-0.5j), 0 \end{matrix} \right). \quad (2.53)$$

Finally, (2.47) is obtained by replacing (2.52) and (2.53). Moreover, in (2.44), for $P_s(e|\gamma) = aQ(\sqrt{2b\gamma})$, we express $Q(\cdot)$ as a Meijer-G function from [39] and [32, 9.31-1] as $aG_{23}^{21} \left(b\gamma \middle| \begin{matrix} 0, 1 \\ 0, 0.5, 0 \end{matrix} \right) / 2\sqrt{\pi}$. Then, (2.44) is written as

$$P_{su}^L = \int_0^\infty P_s(e|\gamma) f_{\gamma_{34}}(\gamma) d\gamma \triangleq \Xi \frac{a}{2\sqrt{\pi}} (P_{s_1}^L - P_{s_2}^L) \quad (2.54)$$

where

$$P_{s_1}^L = \left(\frac{\psi_{41}}{\psi_{34}} \right)^\Lambda \int_0^\infty \gamma^\Lambda G_{23}^{21} \left(b\gamma \middle| \begin{matrix} 0, 1 \\ 0, 0.5, 0 \end{matrix} \right) \times G_{13}^{21} \left(\gamma \psi_{41} \psi_{34} \middle| \begin{matrix} 0 \\ \Lambda, -\Lambda, 0 \end{matrix} \right) d\gamma \quad (2.55)$$

and

$$P_{s_2}^L = \sum_{j=0}^{m_{43}-1} \frac{\beta^j}{j!} \left(\frac{\psi_{43}}{\psi_{34}} \right)^{\Lambda-0.5j} \int_0^\infty \gamma^{\Lambda+0.5j} G_{23}^{21} \left(b\gamma \middle| \begin{matrix} 0, 1 \\ 0, 0.5, 0 \end{matrix} \right) \times G_{13}^{21} \left(\gamma \psi_{43} \psi_{34} \middle| \begin{matrix} 0 \\ \Lambda-j/2, -(\Lambda-j/2), 0 \end{matrix} \right) d\gamma. \quad (2.56)$$

Using [40, 26], (2.55) and (2.56) are expressed as

$$P_{s_1}^L = \left(\frac{\psi_{41}}{\psi_{34}} \right)^\Lambda b^{-(\Lambda+1)} G_{45}^{33} \left(\frac{\psi_{41} \psi_{34}}{b} \middle| \begin{matrix} 0, -\Lambda, -\Lambda-0.5, -\Lambda \\ \Lambda, -\Lambda, -\Lambda, -\Lambda-1, 0 \end{matrix} \right) \quad (2.57)$$

and

$$P_{s_2}^L = \sum_{j=0}^{m_{43}-1} \frac{\beta^j}{j!} \left(\frac{\psi_{43}}{\psi_{34}} \right)^{\Lambda-0.5j} b^{-(\Lambda+0.5j+1)} \times G_{45}^{33} \left(\frac{\psi_{43} \psi_{34}}{b} \middle| \begin{matrix} 0, -(\Lambda+0.5j), -(\Lambda+0.5j)-0.5, -(\Lambda+0.5j) \\ (\Lambda-0.5j), -(\Lambda-0.5j), -(\Lambda+0.5j), -(\Lambda+0.5j+1), 0 \end{matrix} \right) \quad (2.58)$$

Finally, the SU SEP is obtained by replacing (2.57) and (2.58) in (2.54) and the overall BEP for linear EH model is given as $P_{b,su}^L \simeq P_{su}^L/k$.

2.4 Performance Evaluation

In this section, we compare the simulation and theoretical results for the BEPs of PU and SU for different system parameters. Results are provided for both linear and nonlinear EH model of the ST battery. The Monte-Carlo method is applied during computer simulations. It is assumed that CSI is perfectly known at all receiving nodes. Furthermore, we compare our proposed system with the non-cooperative (direct link) system in which PU only benefits the spectrum [36]. Theoretical and simulation results are found in perfect match with each other verifying our derivations.

Unless specifically stated, we assume that $\eta = 0.8$, $\nu = 2.7$ as in [41]. Moreover, 4-QAM modulation is considered for the proposed system while BPSK modulation is assumed for the direct transmission which guarantees equal bandwidth efficiency for both systems with overall transmissions at two and one time slots, respectively. Noise components at all receiving nodes are assumed to have equal variance σ^2 , $P_{min} = 5$ dB and $P_{th} = 25$ dB. Nodes are co-linearly located. We assume that SR is located at the middle of PT and PR where $d_{31} = d_{23} = d_{21}/2$. Moreover, it is assumed that $d_{41} = d_{34} = d_{31}/2$. For all channels, we set the channel powers normalized to $E\{|h|_{ij}^2\} = \Omega_{ij}^2 = 1$. The residual self-interference at SR is set to -30 dB. The power consumed at SR during the two time slots is set to $P_{sr} = 30$ dB where we assume $P_{sr1} = P_{sr2} = 0.5P_{sr}$ for the first and second time slot usages.

Figure 2.2 represents the PU BER versus P_{pt} . BER is provided for two different distances $d_{21} = 1$ and 3. Theoretical curves from (2.10) and (2.20) are provided for the BER of linear and nonlinear EH model at ST battery, respectively. We observe from Figure 2.2 that for a BER value of 10^{-4} , we have approximately SNR gains of 1 dB and 5 dB in comparison with the nonlinear EH model and the direct transmission (non-cooperative), respectively. However, we have an SNR loss for the linear EH model compared to the nonlinear EH model and the direct transmission case. It is observed that by increasing the distance between PT and PR, the nonlinear EH model provides more SNR gain compared to the non-cooperative case. Besides, SU can benefit from the spectrum and transmit its own data to SR.

Figure 2.3 shows the BER performance of SU versus P_{pt} . Here, (2.34) and (2.54) are used for theoretical results. It is shown that for both linear and nonlinear EH models at

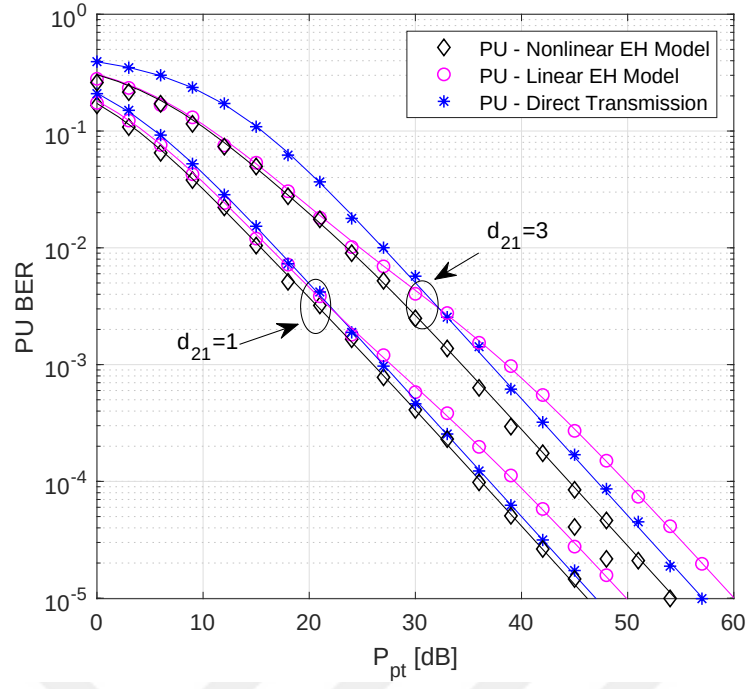


Figure 2.2 : PU BER performance of a wireless-powered CR network over P_{pt} , (lines: theoretical, markers: simulation).

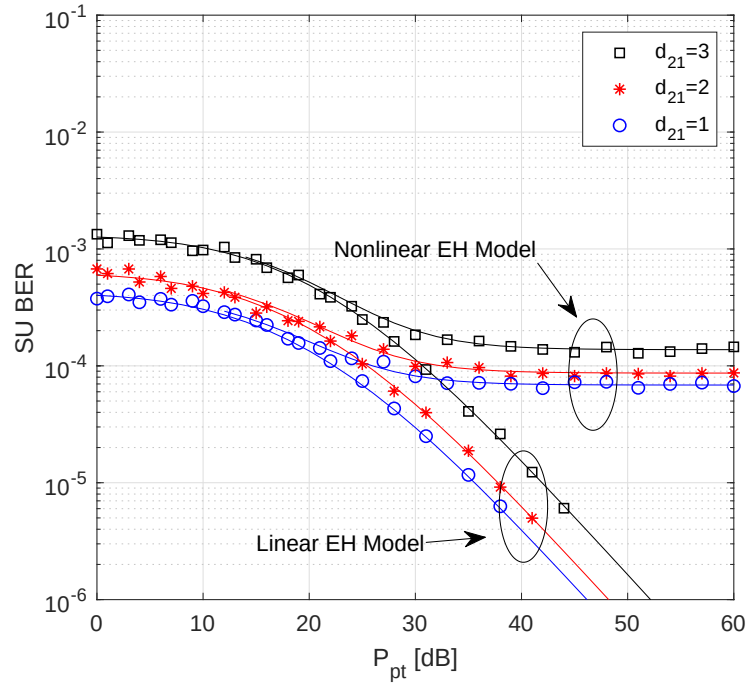


Figure 2.3 : SU BER performance of a wireless-powered CR network over P_{pt} , (lines: theoretical, markers: simulation).

ST, SNR gains can be achieved for smaller distances. Please note that for the nonlinear EH model, we have an error floor because at high SNR region, the harvested power

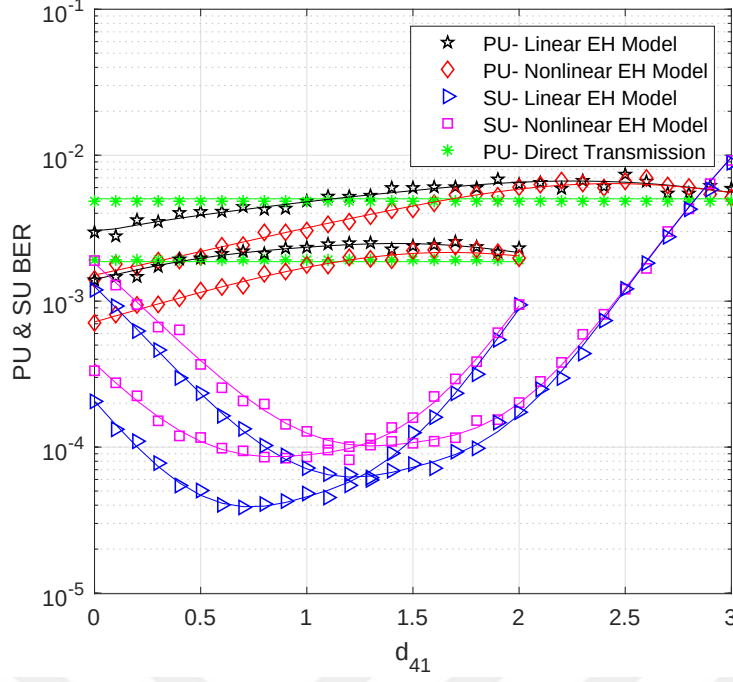


Figure 2.4 : PU & SU BER performance of a wireless-powered CR network over d_{41} where $P_{pt} = 30$ dB, (lines: theoretical, markers: simulation).

reaches to a predefined threshold value which means a constant transmit power from ST to SR.

Figure 2.4 illustrates the PU and SU BERs for the proposed overlay CR system versus the distance between PT and ST. In Figure 2.4, we only move ST node and fix other nodes. We obtain the BER of both PU and SU by moving the node ST co-linearly from PT to PR. Furthermore, as previously stated, we assume $d_{31} = d_{23} = d_{41}/2$. Please note that by increasing or decreasing d_{41} , the distance between links ST→SR, the amount of harvested power from links PT→ST and SR→ST, the effect of interference link from ST to PR changes. Moreover, for realizing the proposed system efficiently, we provide the BER of direct transmission [36] which has a constant BER through the given results. It is observed from Figure 2.4 that when ST is located in the middle of PT and SR, we obtain the best BER result for the proposed system for both linear and nonlinear EH models which is due to the trade-off between harvested energy from PT and caused interference at PR by ST. For example, d_{41} should be smaller than 1.5 and 1 for $d_{21} = 3$ and 2, respectively. The similar trend can be observed for the linear case. For SU, ST harvests RF energy from both SR and PT. Then, to get more received SNR

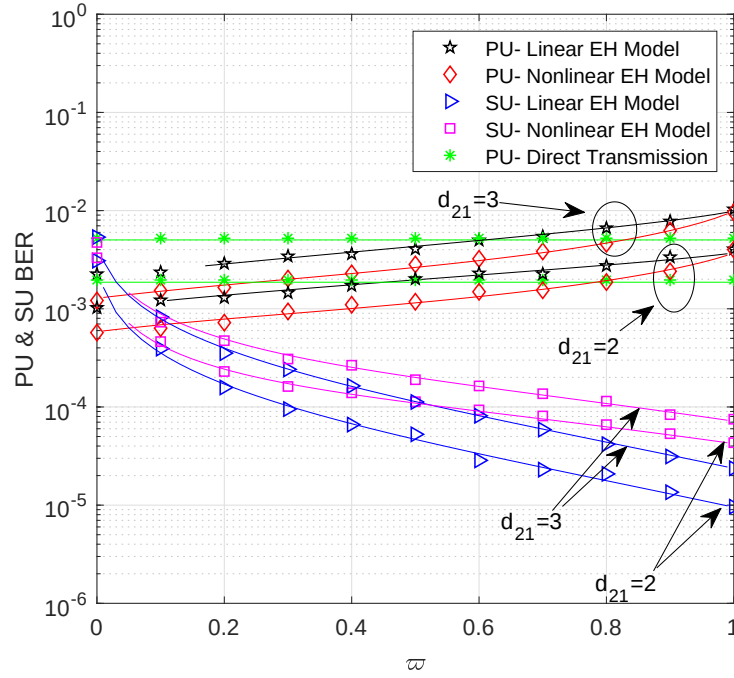


Figure 2.5 : PU & SU BER performance of a wireless-powered CR network over ϖ where $P_{pt} = 30\text{dB}$, (lines: theoretical, markers: simulation).

at SR, ST should be located at a closer distance to PT (between PT and SR) for which P_h gets a high value. Besides, it provides better performance for PU.

Figure 2.5 compares the BER of PU and SU with respect to the ratio ϖ . Here, $P_{sr1} = \varpi P_{sr}$ and $P_{sr2} = (1 - \varpi)P_{sr}$. Moreover, the performance of direct transmission [36] is provided to compare with that of our proposed system. We assume that $P_{pt} = 30\text{ dB}$. Furthermore, results are given for $d_{21} = 2$ and 3 where we assume that SR is located in the middle of PT and PR, and $d_{41} = d_{21}/4$. Based on the proposed system model, P_{sr} is divided into two time slots as P_{sr1} and P_{sr2} , for EH at ST and data transmission to PR, respectively. It is observed from Figure 2.5 that by increasing P_{sr1} (more EH at ST), SU performance improves for both linear and nonlinear EH model; however, for PU, the system performance gets worse due to the strong interference link from ST due to more transmit power. It is shown in Figure 2.5 that for some smaller P_{sr1} values, PU BER performance is better compared to direct link. For example, for $d_{21} = 3$ and 2 , P_{sr1} should be smaller than 0.5 . It is important to note that some theoretical values are not provided in Figure 2.5 due to the properties of Meijer-G function.

Figure 2.6 shows the PU and SU BER performances with respect to P_{th} . Results are given for $P_{pt} = 30\text{ dB}$ and $P_{min} = 5\text{ dB}$. Please note that the results are provided for

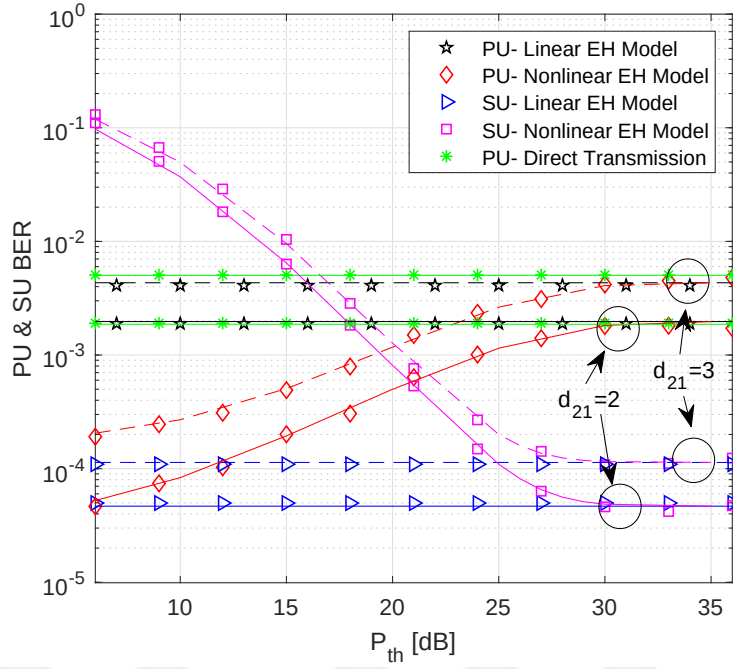


Figure 2.6 : PU & SU BER performance of a wireless-powered CR network over P_{th} where $P_{pt} = 30\text{dB}$ and $P_{min} = 5\text{ dB}$ (lines: theoretical, markers: simulation).

$P_{th} \geq 6\text{ dB}$ since $P_{th} \geq P_{min}$. Moreover, the PU and SU performances are obtained for $d_{21} = 3$ and 2 where $d_{31} = d_{21}/2$ and $d_{41} = d_{31}/2$. It is seen from Figure 2.6 that by increasing P_{th} , the linear and nonlinear EH models provide the same performance. From Figure 2.6, the PU performance is better than that of the direct transmission for $d_{21} = 3$ while for $d_{21} = 2$, we have comparable performances.

The effect of parameter P_{pt} verifies in Figure 2.7 the theoretical derivation of the proposed overlay CR system in Appendix F where P_h is expressed in (2.4) and calculated in (2.9) for both linear and nonlinear EH models. It is shown in Figure 2.7 that the distance d_{21} have an important impact on the amount of harvested energy. For both EH models, smaller distances provide more P_h . It is seen that for the linear EH model, P_h increases by increasing P_{pt} ; However, for the nonlinear EH model, P_h reaches a predefined threshold value for all considered distances as given in (2.9). Please note that for small distance values, P_{st} immediately reaches the P_{th} value.

Figure 2.8 shows P_{st} versus P_{min} . It is seen that by increasing the distances, the transmit power of ST decreases, as expected. Moreover, for $d_{21} = 10$ and $d_{21} = 5$, by increasing P_{min} and taking $P_{th} = 25\text{ dB}$, the average transmit power of ST decreases since the

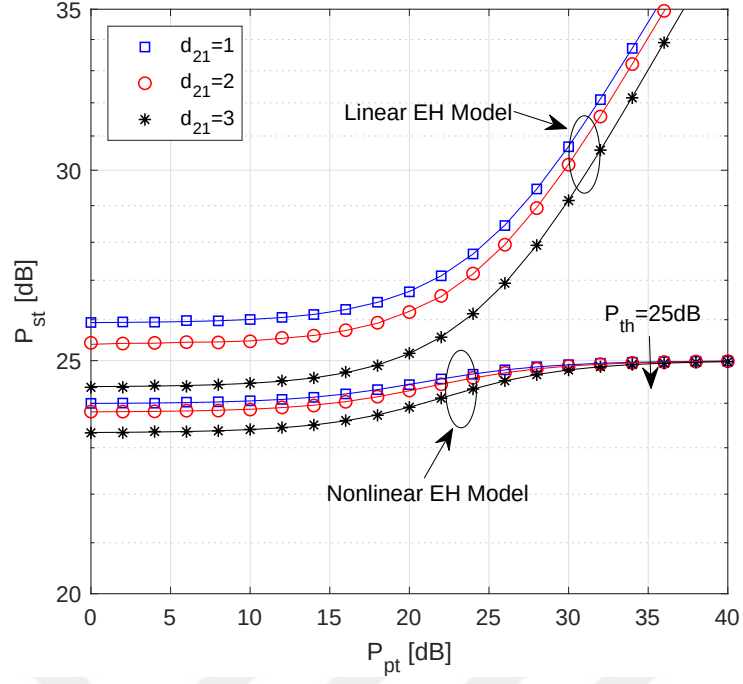


Figure 2.7 : P_{st} of the proposed wireless-powered CR network versus P_{pt} , (lines: theoretical, markers: simulation).

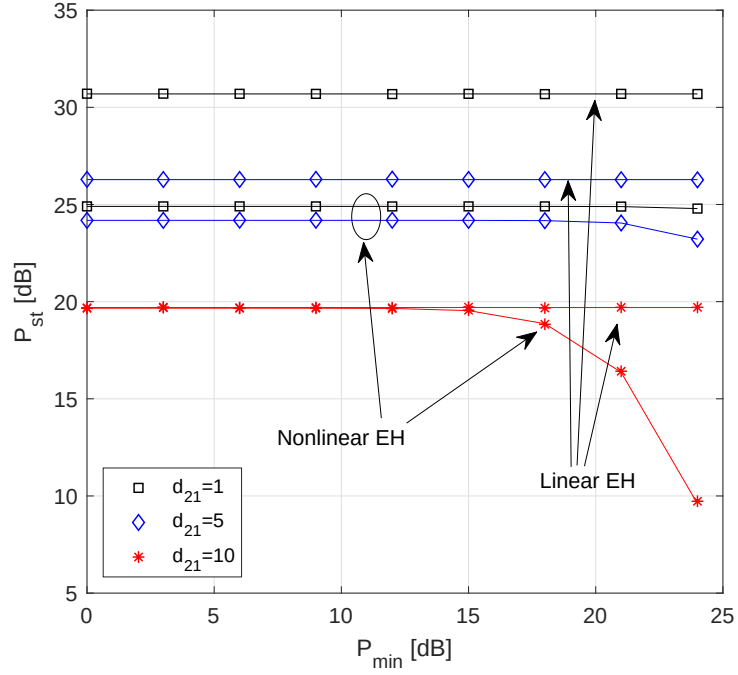


Figure 2.8 : P_{st} of the proposed wireless-powered CR network versus P_{min} , (lines: theoretical, markers: simulations).

average harvested power is smaller than P_{min} . Moreover, for a smaller distance as $d_{21} = 1$ and nonlinear EH model, more average power is harvested and increasing P_{min}

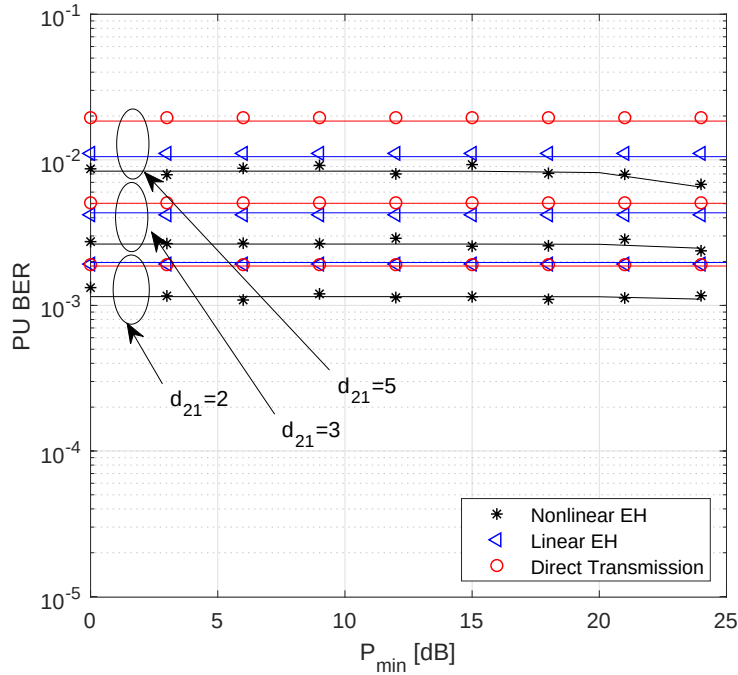


Figure 2.9 : PU BER performance of a wireless-powered CR network versus P_{min} , (lines: theoretical, markers: simulations).

does not affect the average P_{st} . Furthermore, the average transmit power of ST for the linear EH model is always greater than that of the nonlinear EH model since no P_{th} and P_{min} are considered in the linear EH model.

Figure 2.9 compares BER performance of PU versus P_{min} for $d_{21} = 2, 3, 5$ and $P_{th} = 25$ dB. The obtained results for the nonlinear EH PU show better BER performance compared to the direct transmission and the considered linear EH model for all distances. Besides, provided results for the direct transmission and the considered linear EH model are independent of P_{min} and are provided for comparison purposes only. It is shown in Figure 2.9 that for P_{min} values higher than 21 dB and for the distances of $d_{21} = 3$ and $d_{21} = 5$, BER performance relatively gets better since $P_{st} = 0$. In other words, interference link (ST $\rightarrow$ PR) is removed when $P_{st} = 0$ which leads to better PU performance. However, for the smaller distances, since more power is harvested which is above P_{min} , ST always transmits its signals to SR which causes interference at PU.

Figure 2.10 depicts SU BER performance with respect to P_{min} for both linear and nonlinear EH models. Here, when P_{min} is increased, ST is not able to transmit its signal and consequently no error occurs. Therefore, the performance gets better since ST

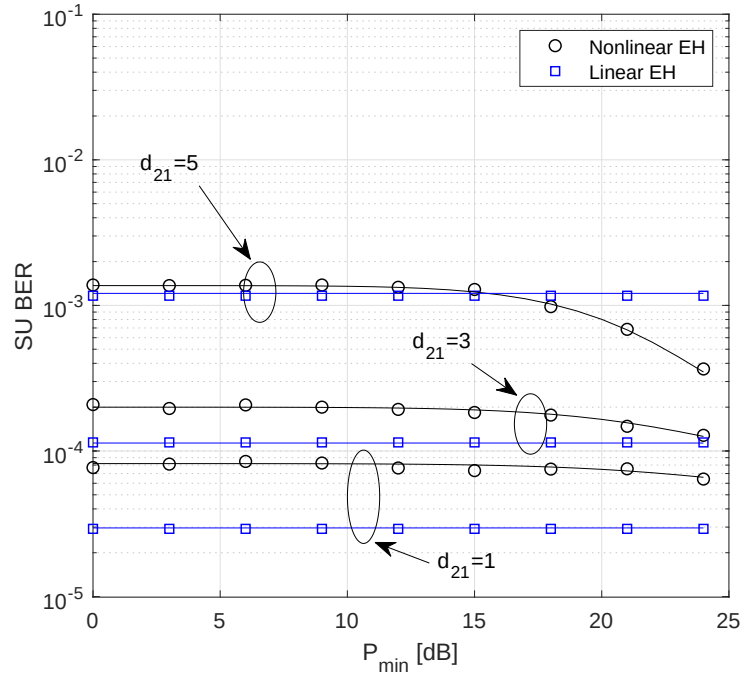


Figure 2.10 : SU BER performance of a wireless-powered CR network versus P_{min} , (lines: theoretical, markers: simulations).

broadcasts its data to SR for high transmit power which guarantees the high received SNR and results in a better BER performance. Also note that by increasing d_{21} and P_{min} , the harvested power gets smaller which leads to no signal transmission from ST.

Figure 2.11 shows the PU BER performance with respect to P_{pt} for 4-QAM and 16-QAM modulation formats. Both theoretical and simulation results are provided for linear and nonlinear EH models. It is seen from Figure 2.11 for both $d_{21} = 3$ and $d_{21} = 2$ that approximately 9 dB and 6 dB SNR losses occur to obtain a BER value of 10^{-4} , for nonlinear and linear EH models, respectively compared to 4-QAM when 16-QAM modulation format is applied.

SU BER performance versus P_{pt} is depicted in Figure 2.12. Results are provided for both 4-QAM and 16-QAM. From Figure 2.12, for a BER value of 10^{-5} for linear EH, 12 dB SNR losses occur, respectively, by increasing modulation order from 4-QAM to 16-QAM, for both $d_{21} = 2$ and $d_{21} = 3$.

Finally, we also analyzed our system under the imperfect CSI. Please note that only the simulation results are provided in Figure 2.13 which shows PU and SU BER

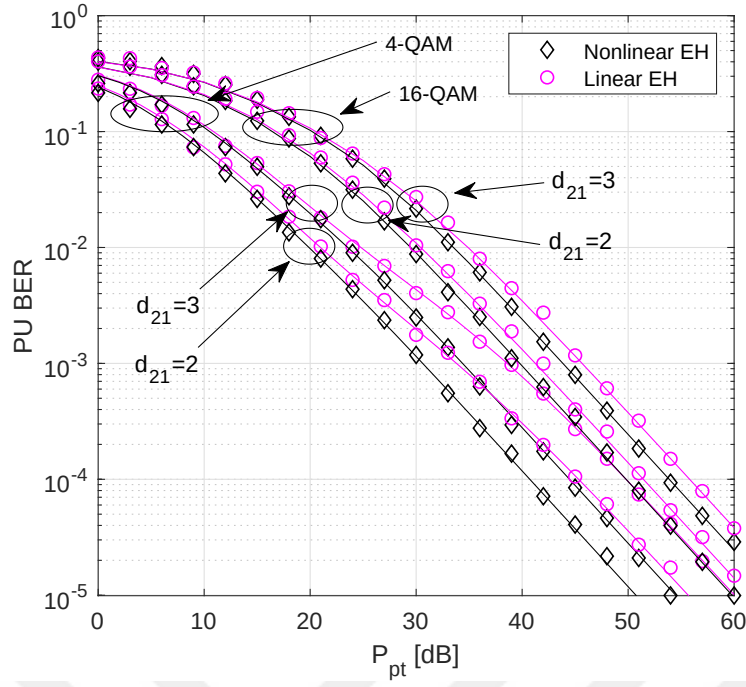


Figure 2.11 : PU BER performance of a wireless-powered CR network versus P_{pt} for $d_{21} = 3$, (lines: theoretical, markers: simulations).

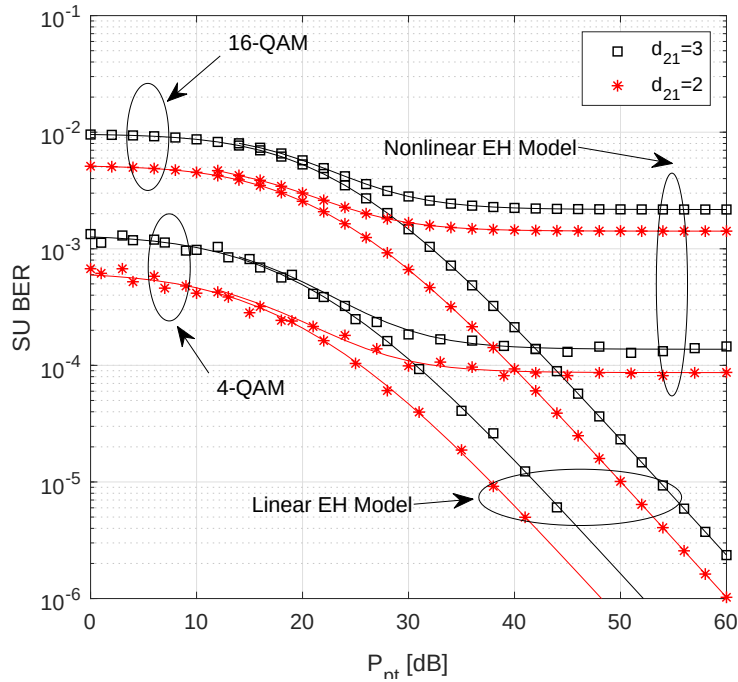


Figure 2.12 : SU BER performance of a wireless-powered CR network versus P_{pt} for $d_{21} = 2$ and $d_{21} = 3$, (lines: theoretical, markers: simulations).

performances versus channel estimation error power σ_{err}^2 . The results are provided for both linear and nonlinear EH models for which the system performances get worse

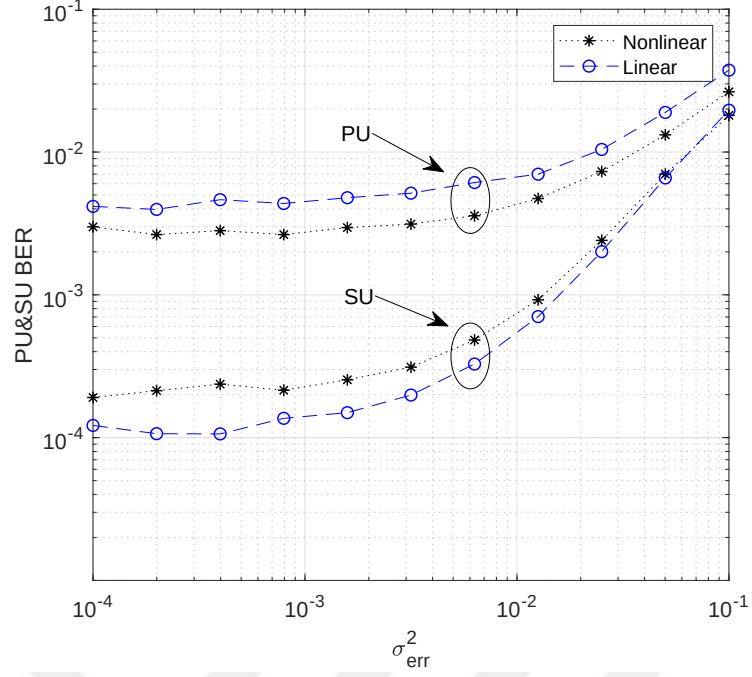


Figure 2.13 : PU and SU BER performance of a wireless-powered CR network versus σ_{err}^2 for $d_{21} = 3$ (simulation results).

when the power of channel estimation error increases.

2.5 Conclusion

In this chapter, we have investigated the BER performance of both PU and SU in the proposed overlay CR network for which overlay spectrum sharing protocol has been considered. ST has been assumed to be a power-constraint node which harvests RF power from both PT and SR while SR operates in FD mode. Moreover, an interference link between ST and PR has been considered which makes the system analysis more realistic. Closed-form expressions have been derived for the BEP of considered overlay CR network and the theoretical results have been validated via computer simulations for different system parameters and for the linear and nonlinear EH models. In addition, the performance of PU in the absence of SU (direct transmission) has been provided for the comparison purposes. The optimum distance value between PT and ST has been obtained which minimizes the BER of both PU and SU. Moreover, the system performances have been obtained with respect to a power allocation coefficient at SR to obtain the best compromise between EH and information relaying.



3. PERFORMANCE ANALYSIS OF DUAL-HOP AF RELAYING WITH NON-LINEAR/LINEAR ENERGY HARVESTING²

Massive device-to-device communication nodes and Internet of Things (IoT) devices are expected to be crucial components in next-generation wireless networks. However, the energy constraint of these nodes presents a challenge since the energy of the batteries is limited. Motivated by this, radio frequency energy harvesting (EH) has been developed as an efficient strategy to overcome the energy constraint of IoT devices and sensor networks. In this chapter, a wireless-powered dual-hop amplify-and-forward relaying system, in the absence of a direct link between the source (S) and the destination (D), is considered. It is assumed that a dedicated power beacon (PB) transmits an energy-bearing signal from which the power-constrained S and relay (R) harvest energy. Theoretical derivations of bit error probability, outage probability, and throughput expressions are performed for both linear and non-linear energy harvesting models. Moreover, the theoretical results provided for different system parameters are validated via Monte Carlo simulations. The obtained results reveal the difference between the realistic non-linear EH model and the conventional linear EH model, which overestimates the system performance at high levels of harvested energy. Thus, it leads to misunderstanding the real performance of the EH systems. However, at low levels of harvested energy, both models behave similarly and provide realistic results.

3.1 Introduction

By deploying massive sensor and Internet of Things (IoT) devices in next-generation wireless communication systems, energy limitation is revealed as one of the challenges for IoT devices. Normally, IoT devices are empowered by a battery, which limits their capacity and constrains their operational time. However, radio frequency (RF) EH can

²This chapter is based on "Babaei, M., Durak-Ata, L. and Aygölü, Ü.(2022). Performance Analysis of Dual-Hop AF Relaying with Nonlinear/Linear Energy Harvesting, Sensors, vol. 22, no. 16: 5987, doi: 10.3390/s22165987."

be an effective solution for empowering IoT devices, leading to increased operational time for battery-free IoT devices. In the literature, simultaneous wireless information and power transfer (SWIPT) and wireless-powered communication (WPC) schemes are considered as RF energy harvesting methods for the power-constrained nodes [4, 14, 21, 34, 42–44]. Power-splitting (PS) and time-switching (TS) are two EH receiver structures utilized in the SWIPT scheme. In PS EH mode, the power-constrained node harvests power from the incoming signal energy where one portion of the signal power is used for harvesting energy while the remainder is used for information processing. Moreover, in TS EH mode, two non-overlapping time intervals are dedicated to EH and IP, respectively. Moreover, the amount of harvested energy is considered as a linear or non-linear function of the energy receiving (ER) node input power [6–10, 14, 21, 22, 31].

3.1.1 Related works

The system performance of power-constrained nodes considering the linear EH model is investigated in [45–52], where the harvested energy is a linear function of the received power at the battery-less node. However, due to the non-linearity of the EH circuit in practice, non-linear EH models are investigated in [53–74].

A massive multiple-input multiple-output (MIMO) relaying system with PS EH mode is investigated in an IoT cooperative network in [45]. The considered system's achievable sum-rate is obtained for a relay which is power-constrained and harvests its power using the linear EH model. The system performance of power-constrained smart devices is investigated in [46], where a TS EH mode with the linear EH model is employed. A full-duplex (FD) dual-hop (DH) relaying network with the linear EH model is studied in [47]. Here, the power-constrained relay node applies PS and TS modes for EH purposes. Moreover, the relay forwards the source node signal to the destination, applying both amplify-and-forward (AF) and decode-and-forward (DF) relaying. Comprehensive throughput analyses of the considered system are performed for both the PS and TS EH modes. The throughput of the considered system is maximized for both PS and TS EH modes. The system performance of a cognitive IoT network with AF/DF relaying is investigated in [48], where the IoT network accesses the spectrum using orthogonal frequency-division multiplexing

(OFDM). Moreover, the linear EH model with PS EH mode is considered in [48]. A cooperative non-orthogonal multiple access (NOMA) overlay spectrum sharing system with power-constrained secondary transmitters (STs) is examined in [49], where the STs operate in FD mode. Moreover, the outage probability and system throughput are investigated considering the linear EH model and PS EH mode. In [50], the outage probability of a cooperative DH relaying system with the linear EH model is derived where the TS EH mode is investigated. Here, the communication between the source and NOMA IoT devices is provided through power-constrained relay nodes.

Maximization of harvested energy in a non-convex problem is investigated in [53]. Here, an unmanned aerial vehicle (UAV) transfers power to the ground user location where the non-linear EH model is considered. The outage probability and throughput of a DH AF relaying two-way (TW) system are investigated in [54], where the relay harvests energy from both source nodes through the TS mode for the non-linear EH model. The throughput and ergodic capacity for a non-linear, piece-wise model are considered in [55], where a power-constrained source node harvests its power from the destination. The performance of a DH AF relaying system with a non-linear EH model is investigated in [56]. A closed-form expression for the bit error probability (BEP) is derived for binary differential phase-shift keying modulation, where the relay is assumed to be power-constrained. The performance of a cooperative AF relaying system is analyzed in [57]. The outage probability is investigated for perfect and imperfect channel state information (CSI), where the relay applies PS for the non-linear EH model. A massive MIMO SWIPT system's energy efficiency is maximized in [58] where a base station transmits power, and the TS EH mode ratio is optimized. The spectral efficiency and average harvested energy of an FD DF relaying network with multiple users are investigated in [59].

The performance of an FD cognitive radio (CR) system for both linear and non-linear EH models is analyzed in [60]. It is assumed that the secondary transmitter is power-constrained and harvests its energy from both the primary transmitter (PT) and the secondary receiver. The bit error rate (BER) performance for both primary and secondary users is derived for different system parameters. The performance of a multi-antenna FD CR EH system is investigated in [61], where the transmitter employs the NOMA technique to transmit from the PT. Both the throughput and the outage

probability of the system based on the non-linear EH model are investigated. In [62], the system performance of a cooperative NOMA network, which consists of three nodes, is analyzed. The near user is considered to be power-constrained and harvests its power from the incoming source signal. The outage probability of an overlay CR is considered in [63]. It is assumed that the secondary user is power-constrained and harvests its energy from the PT signal. The system performance of a DH DF relaying system with multiple power-constrained relays is investigated in [64]. A closed-form outage probability expression is derived considering the TS EH protocol. A cooperative DF relaying network with spectrum sensing and an ER node is examined in [65]. The power-constrained node simultaneously harvests energy and processes information by applying the PS technique.

A MIMO wireless power transfer ER architecture design is studied in [66]. The total harvested power is maximized for the proposed generic architecture of multiple ER nodes. The rate-energy trade-off of a point-to-point SWIPT system with a non-linear EH model is investigated in [67]. The ergodic fading channel is considered in [67] where the power-constrained node applies the PS EH mode. The sum throughput of the relay-based wireless-powered system is investigated in [68]. The relay adopts the non-linear EH model to harvest power from a dedicated power beacon. The sum throughput is maximized jointly considering a non-convex problem with parameters of power and time fraction. A non-linear EH model for a device-to-device network is proposed in [69]. In this study, the energy efficiency is maximized, considering the PS EH mode of the IoT nodes by optimizing the resource and power allocation. A WPC network with a non-linear EH model is investigated in [70], where the weighted sum of computation bits in each user is maximized for the considered non-convex optimization problem. A cooperative DH AF relaying system with a mixed fading environment and a non-linear EH model is investigated in [71]. Here, it is assumed that the relay is power-constrained and harvests its power from the source using PS/TS EH modes. The system performance of a MIMO IoT network in the presence of cooperative jamming is considered in [72], where a non-linear EH model with PS protocol is assumed. In [73], the non-linear EH model is applied to UAV-assisted FD IoT networks, where infinite and finite blocklength codes are considered. The performance of the system is analyzed in terms of the block error rate, where the theoretical derivation is performed

considering Rician shadowed fading channels. Our study can be extended for finite blocklength considering [73] and references therein.

3.2 Methods

In this chapter, the theoretical expressions for the bit error probability (BEP) and the throughput of a wireless-powered dual-hop amplify-and-forward (WP DH AF) relaying system are considered. Moreover, we investigate a non-linear piece-wise EH model which is mathematically tractable in terms of the PDF, CDF of harvested power, and the system performance. Apart from the linear EH model described in the literature, this model provides practical performance analyses. Specifically, the non-linear EH model considered in [8, 31, 56, 57, 64, 75] is addressed since this model is mathematically tractable compared to other non-linear EH models [6, 7, 9, 10]. In contrast to the linear EH model, which overestimates system performance, the considered non-linear piece-wise EH model provides realistic system performance along with the practical non-linear EH models. Since most papers in the literature have considered the linear EH model, in this chapter, the effect of the linear EH model is investigated along with the non-linear piece-wise EH model, to provide a comprehensive system analysis. Hence, comparisons of linear and non-linear EH models are also provided. To the best of the authors' knowledge, the BEP, outage probability, and throughput of the considered system have not yet been investigated in the literature. The contributions of this chapter are summarized as follows:

- Theoretical derivation of BEP expressions considering both linear and non-linear EH models is performed.
- Throughputs of the considered system for both linear and non-linear EH models are derived.
- Simulation and theoretical results of the considered system are provided for different system parameters, such as distance, power, and achievable rate.

The chapter is organized as follows: The system model is described in Section 3.3. Theoretical BEP derivations are provided in Section 3.4. Section 3.5 deals with the

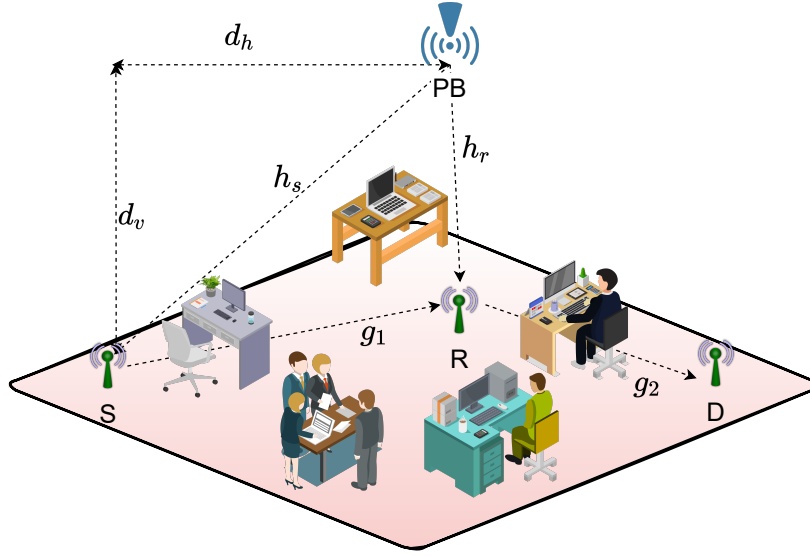


Figure 3.1 : Wireless-powered DH AF relaying system model where S and R harvest their energy from a dedicated PB. S and R use their harvested power for data transmission.

throughput analysis. Numerical results are presented in Section 3.6. Finally, Section 3.7 concludes the chapter.

3.3 System Model

The DH AF relaying system under consideration is given in Figure 3.1, where the source (S) and relay (R) are power-constrained nodes and harvest their energy from a dedicated power beacon (PB). Hence, S and R use their harvested powers for data transmission. In the absence of a direct link between S and the destination (D), the communication is provided with the help of R. All nodes are equipped with only one antenna.

A typical RF EH process is achieved in a short range and a line of sight (LoS) exists between PB and the EH nodes. Inspired by this fact, we assume that the links between nodes in Figure 3.1 are exposed to mix-fading [60, 76]. The proper fading channel model, including an LoS component, is considered as the Rician model. However, the cumulative distribution function (CDF) and probability distribution function (PDF) of the Rician distribution include special functions which make the analysis more complicated and the results are not mathematically tractable. On the other hand, the Nakagami- m fading model provides a good approximation of the Rician channel model [77–79]. Motivated by this, we assume that the links PB→S and PB→R are

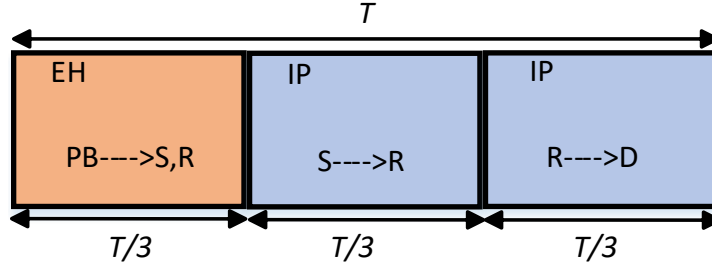


Figure 3.2 : Transmission time frame of the considered wireless-powered system. The transmission frame is assumed to be equally-partitioned into three time slots for EH and information processing (IP).

Table 3.1 : List of notations.

Notation	Description
$\Gamma(\cdot, \cdot)$	Upper incomplete Gamma function ([32, 8.350])
$F(\cdot)$	Cumulative distribution function (CDF)
$f(\cdot)$	Probability distribution function (PDF)
$\Gamma(\cdot)$	Gamma function ([32, 8.310]).
$\gamma(\cdot, \cdot)$	Lower incomplete Gamma function ([32, 8.350])
$(\cdot)!$	Pochhammer's symbol ([80, 6.1-22])
${}_1F_1(\cdot; \cdot; \cdot)$	Kummer Confluent Hyper-geometric function ([38, 07.20.07.0001.01])
$G_{pq}^{mn} \left(\cdot \middle \begin{matrix} a_p \\ b_q \end{matrix} \right)$	Meijer-G function ([32, 9.301])
$K_x(\cdot)$	Modified Bessel function of the second kind with order x [32]

exposed to Nakagami- m fading, represented by channel coefficients h_s and h_r , with channel parameters m_z and m_w , respectively. Moreover, the links $S \rightarrow R$ and $R \rightarrow D$ are subject to Rayleigh fading, with gains g_1 and g_2 , respectively. It is assumed that the CSI is perfectly known at the receiving nodes, and the channels are assumed to be exposed to flat fading and remain fixed during a transmission interval and vary independently from one interval to another. The overall transmission time T is divided into three equal time slots, as shown in Figure 3.2. It should be noted that this assumption is considered in [14, 21] where it provides a maximized system performance. In the first time slot of duration $T/3$, PB broadcasts the dedicated energy bearing signal. In the second and third time slots, S and R transmit their signals to R and D, respectively. n_r and n_d denote additive white Gaussian noise (AWGN) samples which are independent and identically distributed (i.i.d.) complex Gaussian random variables (r.v) as $n_r, n_d \sim \mathcal{CN}(0, N_0)$. Finally, the notation and system parameters are listed in Tables 3.1 and 3.2, respectively.

Table 3.2 : List of parameters.

Parameters	Description
P_T	PB transmit power
η_s	Energy conversion coefficient ($0 < \eta_s < 1$)
η_r	Energy conversion coefficient ($0 < \eta_r < 1$)
P_s	Harvested power at node S
P_r	Harvested power at node R
x	Transmitted signal from S with $E\{ x ^2\} = 1$
a, b	Modulation specific constants [21]
$\Omega_{h_i} = E\{ h_i ^2\}$	Channel fading gains, $i \in \{s, r\}$
$\Omega_{g_j} = E\{ g_j ^2\}$	Channel fading gains, $j \in \{1, 2\}$
P_{th_i}	Threshold (saturation) power at nodes source and relay where $i \in \{s, r\}$
P_{th_i}	Threshold (saturation) power at source and relay nodes $i \in \{s, r\}$
ν	Path loss exponent
$L_i = 1/\sqrt{1 + d_i^\nu}$	Pathloss $i \in \{st, rt, sr, rd\}$ [21, 34]
d_{sr}	S→R link distance
d_{rd}	R→D link distance
d_{st}	Distances from PB to S
d_{rt}	Distances from PB to R
d_v	Vertical distance from PB to node S
d_h	Horizontal distance from S to PB
N_0	Noise power spectral density
$\vartheta = \log_2 M$	Number of bits per symbol
M	Modulation order

3.3.1 Linear and non-linear EH models

In the literature, two EH models are considered: linear and non-linear. In the linear EH model, the transmit power of the power-constrained node is increased by increasing the harvested power at the considered node [14, 21]. This causes a misrepresentation of the amount of harvested power [1] since, in practical EH circuits, the amount of the harvested power increases to a threshold level rather than the considered amount of harvested power in the linear EH model.

In other words, in practice, due to the non-linear behavior of the diodes in EH circuits, non-linear EH models are more realistic compared to the linear EH model. Additionally, a maximum threshold power is defined for the non-linear EH model, such that, for harvested power greater than this threshold value, the transmit powers of S and R take this fixed threshold value [6–10, 31]. Specifically, we assume the

non-linear EH model given in [31] and [8] since the non-linear EH models proposed in [6, 7, 9, 10] are not mathematically tractable. Moreover, the considered model provides a good approximation of practical EH circuits at low and high amounts of harvested power [1, 31] which broadens the insight of EH system design along with the well-studied linear EH model.

In the first time slot, S and R harvest energy from PB. For the linear EH model, the harvested power for S and R is given as [14]

$$P_i = P_{h_i} = \eta_i P_T L_{it} |h_i|^2, \quad (3.1)$$

where P_{h_i} is the harvested power at node i with $i \in \{s, r\}$. However, assuming the non-linear EH model, the harvested power at S and R is given as [8, 31]

$$P_i = \begin{cases} P_{h_i}, & P_{h_i} \leq P_{th_i} \\ P_{th_i}, & P_{h_i} > P_{th_i} \end{cases} \quad (3.2)$$

where $i \in \{s, r\}$. We assume that all of the harvested energy at both nodes S and R during the first time slot of $T/3$ is consumed for the transmission of the signal x in the consecutive time slots of each $T/3$ since there is no available energy buffer at the power-constrained nodes. Moreover, high input power is limited to the threshold power P_{th_i} . In the second time slot, the received signal at node R is

$$y_r = \sqrt{P_s L_{sr}} g_1 x + n_r. \quad (3.3)$$

In the third time slot, R forwards the amplified version of the received signal in (3.3). The received signal at D is then given as

$$y_d = \sqrt{\frac{P_r L_{rd}}{G}} g_2 y_r + n_d, \quad (3.4)$$

where $G = E\{|y_r|^2\} = P_s L_{sr} |g_1|^2 + N_0$ is the normalization factor. Here, it is assumed that the variable-gain AF relaying protocol is considered. Substituting (3.3) in (3.4), we have

$$y_d = \sqrt{\frac{P_s P_r L_{sr} L_{rd}}{G}} g_1 g_2 x + \sqrt{\frac{P_r L_{rd}}{G}} g_2 n_r + n_d. \quad (3.5)$$

The received SNR at D is calculated from (3.5) as

$$\gamma = \frac{XY}{X+Y+1} \leq \min(X, Y), \quad (3.6)$$

where $X = ZX'$, $Y = WY'$ and we assume a tight upper bound in (3.6). For simplicity, we assume $Z = P_s$, $W = P_r$, $X' = L_{sr}|g_1|^2/N_0$ and $Y' = L_{rd}|g_2|^2/N_0$. In addition, $Z \sim \Gamma(m_z, \phi_z)$ and $W \sim \Gamma(m_w, \phi_w)$ are Gamma distributed r.v.s with parameters m_z and m_w , respectively, and the scaling parameters are $\phi_z = m_z/\bar{z}$ and $\phi_w = m_w/\bar{w}$. Moreover, $\bar{z} = \eta_s P_T L_{st} \Omega_{h_s}$ and $\bar{w} = \eta_r P_T L_{rt} \Omega_{h_r}$. Furthermore, X' and Y' are exponentially distributed as $\text{Exp}(\bar{x}')$ and $\text{Exp}(\bar{y}')$, where $\bar{x}' = L_{sr} \Omega_{g_1}/N_0$ and $\bar{y}' = L_{rd} \Omega_{g_2}/N_0$, respectively.

In (3.6), using [37, eq. 6-81], the conditioned CDF of γ is expressed as

$$F_\gamma(\gamma|Z, W) = F_X(\gamma|Z) + F_Y(\gamma|W) - F_X(\gamma|Z)F_Y(\gamma|W), \quad (3.7)$$

where

$$F_X(\gamma|Z) = 1 - \exp\left(-\frac{\gamma}{Z\bar{x}'}\right), \quad (3.8)$$

$$F_Y(\gamma|W) = 1 - \exp\left(-\frac{\gamma}{W\bar{y}'}\right). \quad (3.9)$$

Substituting both (3.8) and (3.9) in (3.7) and simplifying, the conditioned CDF in (3.7) is obtained as

$$F_\gamma(\gamma|Z, W) = 1 - \exp\left(-\gamma\left(\frac{1}{Z\bar{x}'} + \frac{1}{W\bar{y}'}\right)\right). \quad (3.10)$$

3.4 BEP Analysis

In this section, the BEP of the considered DH AF relaying system is derived for both non-linear and linear EH models. In order to obtain the BEP expression, the analytical expression of symbol error probability (SEP) is first derived. The overall BEP expression is given for high SNR, assuming the common approximation for Gray mapping [36] as $P_b^i \cong P_s^i / \vartheta$. Here, $i \in \{L, NL\}$, L and NL represent linear and non-linear EH models, respectively. The conditioned SEP for both linear and non-linear EH models is calculated as [21]

$$P_s^{L,NL}(e|Z, W) = \frac{a\sqrt{b}}{2\sqrt{\pi}} \int_0^\infty \frac{e^{-b\gamma}}{\sqrt{\gamma}} F_\gamma(\gamma|Z, W) d\gamma = \frac{a\sqrt{b}}{2\sqrt{\pi}} (I_1 - I_2(Z, W)), \quad (3.11)$$

where parameters a and b denote the modulation coefficients for M -PSK/QAM [21] and $F_\gamma(\gamma|Z, W)$ is given in (3.10). In (3.11),

$$I_1 = \int_0^\infty \frac{e^{-b\gamma}}{\sqrt{\gamma}} d\gamma = \sqrt{\frac{\pi}{b}} \quad (3.12)$$

and

$$I_2(Z, W) = \int_0^\infty \frac{1}{\sqrt{\gamma}} \exp\left(-\gamma\left(b + \frac{1}{Z\bar{x}'} + \frac{1}{W\bar{y}'}\right)\right) d\gamma = \sqrt{\frac{\pi}{b + \frac{1}{Z\bar{x}'} + \frac{1}{W\bar{y}'}}} \quad (3.13)$$

which are calculated using [32, eq. 3.361-2]. After substituting (3.12) and (3.13) and simplifying, (3.11) is obtained as

$$P_s^{L,NL}(e|Z, W) = \frac{a}{2} \left[1 - \frac{1}{\sqrt{1 + \frac{1}{Z\bar{x}} + \frac{1}{W\bar{y}}}} \right], \quad (3.14)$$

where $\bar{x} = b\bar{x}'$ and $\bar{y} = b\bar{y}'$.

3.4.1 BEP analysis of the non-linear EH model

In this subsection, the BEP of the non-linear EH model is derived analytically. The overall BEP for the non-linear EH model is expressed as

$$P_b^{NL} \cong (P_{s1} + P_{s2} + P_{s3} + P_{s4}) / \vartheta, \quad (3.15)$$

where P_{s1} , P_{s2} , P_{s3} , and P_{s4} stand for the four states of power harvesting processes considering both nodes S and R, which are calculated as

$$P_{s1} = \int_{w=0}^{P_{th}} \underbrace{\int_{z=0}^{P_{th}} P_s^{L,NL}(e|z, w) f_Z(z) dz}_{A(w)} f_W(w) dw, \quad (3.16)$$

$$P_{s2} = P_s^{L,NL}(e|_{w=P_{th}}^{z=P_{th}}) \underbrace{\int_{w=P_{th}}^{\infty} f_W(w) dw}_{B_1} \underbrace{\int_{z=P_{th}}^{\infty} f_Z(z) dz}_{B_2}, \quad (3.17)$$

$$P_{s3} = \underbrace{\int_{z=0}^{P_{th}} P_s^{L,NL}(e|z, w) f_Z(z) dz}_{A_1=A(w)|_{w=P_{th}}} \underbrace{\int_{w=P_{th}}^{\infty} f_W(w) dw}_{B_1}, \quad (3.18)$$

and

$$P_{s4} = \underbrace{\int_{w=0}^{P_{th}} P_s^{L,NL}(e|z, w) f_W(w) dw}_{A_2=A(z)|_{z=P_{th}}} \underbrace{\int_{z=P_{th}}^{\infty} f_Z(z) dz}_{B_2}. \quad (3.19)$$

$A(w)$, and P_{s1} in (3.16) and $A(z)$ in (3.19) are calculated in Appendix G, Appendix H, and Appendix I, respectively.

In (3.19), A_2 is calculated from (I.2) as $A_2 = A(z)|_{z=P_{th}}$. Furthermore, using [32, eq. 3.351-2], B_2 in (3.19) is calculated as

$$B_2 = \int_{P_{th}}^{\infty} f_Z(z) dz = D \Gamma(m_z, P_{th} \phi_z) / \phi_z^{m_z}, \quad (3.20)$$

where

$$f_Z(z) = D z^{m_z-1} \exp(-z \phi_z) \quad (3.21)$$

with $D = \phi_z^{m_z} / \Gamma(m_z)$. Finally, P_{s4} in (3.19) is calculated by substituting $A_2 = A(z)|_{z=P_{th}}$ in (I.2) and (3.20).

In (3.18), A_1 is calculated from (G.2) as $A_1 = A(w)|_{w=P_{th}}$. Moreover, using [32, 3.351-2], B_1 in (3.18) is calculated as

$$B_1 = \int_{P_{th}}^{\infty} f_W(w) dw = I \Gamma(m_w, P_{th} \phi_w) / \phi_w^{m_w}, \quad (3.22)$$

where

$$f_W(w) = I w^{m_w-1} \exp(-w \phi_w) \quad (3.23)$$

with $I = \phi_w^{m_w} / \Gamma(m_w)$. Moreover, P_{s3} in (3.18) is calculated by substituting $A_1 = A(w)|_{w=P_{th}}$ in (G.2) and (3.22). Finally, P_{s2} in (3.17) is calculated using (3.22), (3.20), and replacing $z = P_{th}$ and $w = P_{th}$ in (3.14).

3.4.2 Linear EH model

In this subsection, the BEP of the considered DH AF relaying system is derived for the linear EH. Here, it is assumed that the harvested energy is linearly dependent on the received energy and increases by increasing the energy transferred from PB. At high SNR values, the overall BEP for the linear EH model is given as

$$P_b^L \cong \frac{1}{\vartheta} \int_0^\infty \underbrace{\int_0^\infty P_s^{L,NL}(e|z, w) f_Z(z) dz}_{P_s(e|w)} f_W(w) dw. \quad (3.24)$$

In (3.24), $P_s(e|w)$ is calculated by substituting $f_Z(z)$ and $P_s^{L,NL}(e|z, w)$ from (3.21) and (3.14) as

$$P_s(e|w) = \frac{a}{2}(E_0 - F_0(w)), \quad (3.25)$$

where

$$E_0 = \int_0^\infty f_Z(z) dz = 1, \quad (3.26)$$

and after simplifying

$$F_0(w) = \frac{D}{\sqrt{B}} \int_0^\infty \frac{z^{m_z-0.5}}{\sqrt{1+z/G}} \exp(-z\phi_z) dz, \quad (3.27)$$

where $G = B/C$. Using [40, eq. 10], (3.27), is rewritten as

$$F_0(w) = \frac{D}{\Gamma(0.5)\sqrt{B}} \int_0^\infty z^{m_z-0.5} G_{11}^{11} \left(\frac{Z}{G} \middle| \begin{matrix} 0.5 \\ 0 \end{matrix} \right) \exp(-z\phi_z) dz. \quad (3.28)$$

Using [32, eq. 7.813-1], and substituting $G = B/C$, (3.28) is obtained as

$$F_0(w) = \frac{D}{\Gamma(0.5)\sqrt{B}} \phi_z^{-(m_z+0.5)} G_{21}^{12} \left(\frac{1+w\bar{y}}{B\phi_z w \bar{y}} \middle| \begin{matrix} 0.5 - m_z, 0.5 \\ 0 \end{matrix} \right). \quad (3.29)$$

Then, (3.25) is calculated by substituting (3.26) and (3.29). Moreover, substituting (3.25) and (3.23) in (3.24) we have

$$P_b^L \cong \frac{1}{\vartheta} \frac{a}{2} \int_0^\infty (1 - F_0(w)) f_W(w) dw = \frac{a}{2\vartheta} (P - J), \quad (3.30)$$

where

$$P = \int_0^\infty f_W(w) dw = 1 \quad (3.31)$$

and

$$J = \frac{DI}{\Gamma(0.5)\sqrt{B}} \phi_z^{-(m_z+0.5)} \int_0^\infty w^{m_w-1} \exp(-w\phi_w) G_{21}^{12} \left(\frac{1+w\bar{y}}{B\phi_z w \bar{y}} \middle| \begin{matrix} 0.5-m_z, 0.5 \\ 0 \end{matrix} \right) dw. \quad (3.32)$$

Please note that no closed-form solution is available for (3.32) as it is calculated numerically. Finally, the overall BEP for the linear EH model is obtained from (3.30) which depends on only J in (3.32).

3.5 Throughput Analysis

In this section, the throughput of the DH AF relaying system is calculated for the non-linear and linear EH models. The outage probability, defined as the probability that the target rate exceeds the instantaneous achievable rate, can be given as

$$P_{out}^i = \mathcal{P}_r \left\{ \frac{1}{3} \log_2(1 + \gamma) < R \right\} = \mathcal{P}_r \{ \gamma < \gamma_{th} \} = F_\gamma(\gamma_{th}), \quad (3.33)$$

where R is the target rate, $\gamma_{th} = 2^{3R} - 1$ is the threshold SNR, γ is the instantaneous SNR at D and $F_\gamma(\cdot)$ is the CDF of γ . The factor $1/3$ is due to the transmission of one symbol per three time slots. Please note that $i = \text{NL}$ and $i = \text{L}$ stand for non-linear and linear EH models, respectively. The throughput of the considered system is calculated as

$$\tau^i = R(1 - P_{out}^i)/3. \quad (3.34)$$

3.5.1 Outage probability for the non-linear EH model

The system outage probability for the non-linear EH model is calculated as

$$P_{out}^{NL} = P_{out_1} + P_{out_2} + P_{out_3} + P_{out_4}, \quad (3.35)$$

where the four outage probabilities at the right-hand side are calculated as

$$P_{out_1} = \int_{w=0}^{P_{th}} \underbrace{\int_{z=0}^{P_{th}} F_{\gamma}(\gamma|z, w) f_Z(z) dz}_{Q(w)} f_W(w) dw, \quad (3.36)$$

$$P_{out_2} = F_{\gamma}(\gamma|_{w=P_{th}}^{z=P_{th}}) \underbrace{\int_{w=P_{th}}^{\infty} f_W(w) dw}_{B_1} \underbrace{\int_{z=P_{th}}^{\infty} f_Z(z) dz}_{B_2}, \quad (3.37)$$

$$P_{out_3} = \underbrace{\int_{z=0}^{P_{th}} F_{\gamma}(\gamma|z, w) f_Z(z) dz}_{Q_1=Q(w)|_{w=P_{th}}} \underbrace{\int_{w=P_{th}}^{\infty} f_W(w) dw}_{B_1}, \quad (3.38)$$

and

$$P_{out_4} = \underbrace{\int_{w=0}^{P_{th}} F_{\gamma}(\gamma|z, w) f_W(w) dw}_{Q_2=Q(z)|_{z=P_{th}}} \underbrace{\int_{z=P_{th}}^{\infty} f_Z(z) dz}_{B_2}. \quad (3.39)$$

In (3.36), substituting $f_Z(z)$ and $f_W(w)$ from (3.21) and (3.23), respectively, and considering $F_{\gamma}(\gamma|z, w)$ in (3.10), we have

$$Q(w) = E_1 - \exp(-\frac{\gamma}{w\bar{y}'}) O_1(\gamma) \quad (3.40)$$

and

$$P_{out_1} = E_1 E_2 - O_1(\gamma) O_2(\gamma). \quad (3.41)$$

Here, E_1 and E_2 are calculated from (G.3) and (I.3), respectively, and

$$O_1(\gamma) = \int_0^{P_{th}} \exp(-\frac{\gamma}{z\bar{x}}) f_Z(z) dz \quad (3.42)$$

and

$$O_2(\gamma) = \int_0^{P_{th}} \exp(-\frac{\gamma}{w\bar{y}'}) f_W(w) dw. \quad (3.43)$$

Substituting, $f_Z(z)$ and $f_W(w)$ from (3.21) and (3.23) in (3.42) and (3.43), respectively, and using [32, eq. 1.211-1], [40, eq. 11] and applying [32, eq. 9.31-1] and [32, eq. 9.31-2] for both (3.42) and (3.43), after simplification, using [40, eq. 26], we have

$$O_1(\gamma) = D \sum_{t=0}^{\Upsilon} \frac{(-\phi_z)^t}{t!} P_{th}^{m_z+t} G_{32}^{12} \left(\frac{P_{th}\bar{x}'}{\gamma} \middle| 1, 1 - (m_z + t), 0 \right) \quad (3.44)$$

and

$$O_2(\gamma) = I \sum_{t=0}^{\Upsilon} \frac{(-\phi_w)^t}{t!} P_{th}^{m_w+t} G_{32}^{12} \left(\frac{P_{th}\bar{y}'}{\gamma} \middle| 1, 1 - (m_w + t), 0 \right). \quad (3.45)$$

In (3.44) and (3.45), Υ is a parameter determining the trade-off between complexity and accuracy. Finally, substituting (3.44) and (3.45) in (3.41), P_{out_1} in (3.41) is derived. In order to calculate (3.37), we replace $z = w = P_{th}$ in $F_\gamma(\gamma|z, w)$ given in (3.10) and calculate B_1 and B_2 from (3.22) and (3.20), respectively. Moreover, for (3.38), Q_1 and B_1 are calculated by replacing $w = P_{th}$ in (3.40) and (3.22), respectively. Pursuing the same procedure for (3.40), (3.39) is expressed as

$$Q(z) = E_2 - \exp\left(-\frac{\gamma}{z\bar{x}'}\right)O_2(\gamma). \quad (3.46)$$

Then, in (3.39), Q_2 and B_2 are calculated by replacing $z = P_{th}$ in (3.46) and (3.20), respectively. Finally, (3.35) is calculated by substituting (3.36), (3.37), (3.38) and (3.39). The throughput of the proposed system is derived by substituting (3.35) in (3.34).

3.5.2 Outage probability for the linear EH model

The outage probability of the linear EH model is calculated as

$$P_{out}^L = \int_0^\infty F_\gamma(\gamma|W)f_W(w)dw, \quad (3.47)$$

where

$$F_\gamma(\gamma|W) = \int_0^\infty F_\gamma(\gamma|Z, W)f_Z(z)dz \quad (3.48)$$

and $F_\gamma(\gamma|Z, W)$ is given in (3.10). Substituting (3.21) and (3.10) and after simplification, (3.48) is obtained as

$$F_\gamma(\gamma|W) = S_{11} - S_{12}, \quad (3.49)$$

where

$$S_{11} = \int_0^\infty f_Z(z)dz = 1 \quad (3.50)$$

and

$$\begin{aligned} S_{12} &= D \exp\left(-\frac{\gamma}{w\bar{y}'}\right) \int_0^\infty z^{m_z-1} \exp\left(-z\phi_z - \frac{\gamma}{z\bar{x}'}\right) dz \\ &= 2D \exp\left(-\frac{\gamma}{w\bar{y}'}\right) \left(\frac{\gamma}{\bar{x}'\phi_z}\right)^{0.5m_z} K_{m_z}\left(2\sqrt{\frac{\gamma\phi_z}{\bar{x}'}}\right). \end{aligned} \quad (3.51)$$

Here, (3.51) is obtained by using [32, eq. 3.471-9]. Substituting (3.49) in (3.47) and simplifying, we have

$$P_{out}^L = S_{21} - S_{22}, \quad (3.52)$$

where

$$S_{21} = S_{11} \int_0^\infty f_W(w) dw = S_{11} = 1 \quad (3.53)$$

and

$$S_{22} = 2DI \left(\frac{\gamma}{\bar{x}'\phi_z} \right)^{0.5m_z} K_{m_z} \left(2\sqrt{\frac{\gamma\phi_z}{\bar{x}'}} \right) \int_0^\infty w^{m_w-1} \exp\left(-w\phi_w - \frac{\gamma}{w\bar{y}'}\right) dw. \quad (3.54)$$

Using [32, eq. 3.471-9], after simplification, the CDF of the SNR at node D, namely, the outage probability is calculated as

$$P_{out}^L = 1 - 4DI \left(\frac{1}{\bar{x}'\phi_z} \right)^{0.5m_z} \left(\frac{1}{\bar{y}'\phi_w} \right)^{0.5m_w} \gamma^{0.5(m_z+m_w)} \\ \times K_{m_z} \left(2\sqrt{\frac{\gamma\phi_z}{\bar{x}'}} \right) K_{m_w} \left(2\sqrt{\frac{\gamma\phi_w}{\bar{y}'}} \right). \quad (3.55)$$

Finally, the throughput of the linear EH model is obtained by substituting (3.55) in (3.34).

3.6 Results and Discussion

In this section, the simulation and theoretical results are presented for different system parameters, which provides a comprehensive insight into the analysis of the WP DH AF relaying system. In all figures, the simulation and theoretical curves are denoted by symbols and lines, respectively. The numerical results obtained from simulations and theoretical derivations are in perfect match with each other at high SNR values, which verifies our theoretical analysis. Moreover, throughout the chapter, the integrals in (3.32), and (H.13) are numerically calculated using MATLAB and Wolfram Mathematica Computer Software. Furthermore, the numerical results for the derived summations in analytical expressions are obtained for $\Upsilon = 200$, which makes them sufficiently reliable. The BER results are provided with respect to various system parameters. The numerical results for BEP are obtained from (3.15) and (3.30) and for throughput by replacing (3.35) and (3.55) in (3.34), for the linear and non-linear EH models, respectively. Unless otherwise stated, we assume $d_v = 1$ and $\eta_s = \eta_r = 0.9$.

The path loss coefficient is taken as $\nu = 2.7$ [41]. All channel gains and noise powers are fixed as $\Omega = 1$ and $N_0 = 1$, respectively. Moreover, R is located at the middle of S and D, so that $d_{sr} = d_{rd} = d_{sd}/2$. It is assumed that S, R, and D are located co-linearly, and PB moves along a trajectory parallel to S-R-D. Then, $d_{st} = \sqrt{d_v^2 + d_h^2}$ and $d_{rt} = \sqrt{d_v^2 + (d_{sr} - d_h)^2}$. Please note that the above assumptions are for numerical examples; the derived analytical expressions are obtained for the general case, and the nodes can be randomly located. The channel parameters for the links S→PB and R→PB are chosen as $m_z = 2$ and $m_w = 2$, respectively. For the BEP analysis, we provide the results for 4-QAM modulation, and for rates $R = 1, 2, 3$ bit/sec/Hz, which are valid for low energy power-constraint nodes. Please note that the approximation in (3.6) provides a tight upper bound at medium and high SNR values for the following BER results.

Figure 3.3 represents the relative approximation error of the throughput with respect to the parameter Υ . Specifically, we define

$$\varepsilon = \left| \frac{\text{analytical value} - \text{simulation value}}{\text{simulation value}} \right| \quad (3.56)$$

as the amount of relative error. Here the simulation value is obtained by the Monte Carlo method while the analytical value is calculated from (3.35). As seen from Figure 3.3, for both $R = 1$ and $R = 3$ bits/sec/Hz, by increasing the value of Υ , the error is decreased and tends to zero. Moreover, the results are sufficiently accurate when they are obtained for $\Upsilon = 7$, which provides an error value of $\varepsilon = 0.0002$. However, we guaranteed the results by taking $\Upsilon = 200$. In addition, the BER results are depicted in Figure 3.4 for different values of Υ considering (3.15). From Figure 3.4, it is concluded that $\Upsilon = 10$ provides more accurate upper bound values compared to $\Upsilon = 1$.

The BER performance of the proposed WP DH AF relaying system with respect to P_T is presented in Figure 3.5. For P_T values lower than 40 dB, the linear and non-linear models provide the same performance, while for P_T higher than 40 dB, the BER performance for the non-linear EH model converges to the error floor, which verifies that the harvested power is higher than the threshold power. In other words, for higher SNR values, $P_s = P_r = P_{th}$. It is seen from Figure 3.5 that the linear EH model overestimates the system performance compared to the realistic results obtained for the non-linear EH model. Moreover, considering the linear EH model, approximately,

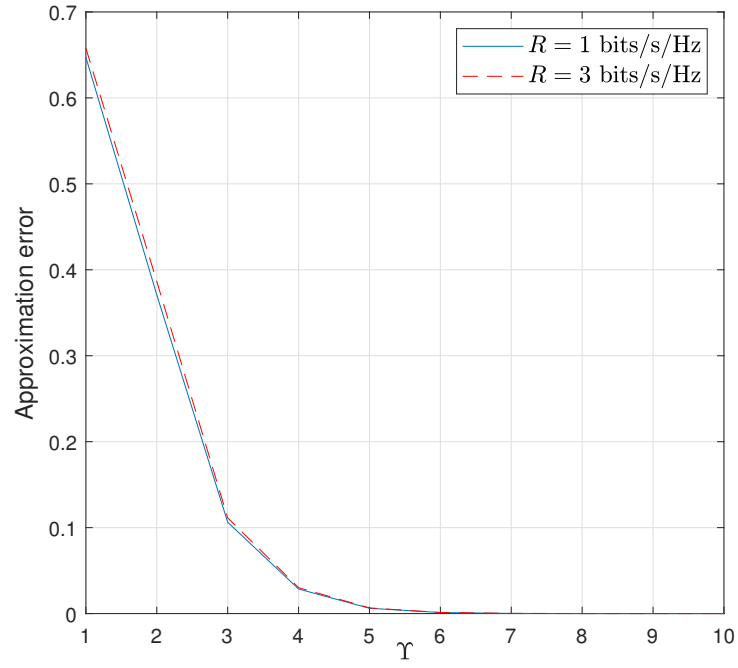


Figure 3.3 : Relative approximation error versus parameter Υ of the WP DH AF relaying system for $P_T = 40$ dB, and $P_{th} = 35$ dB.

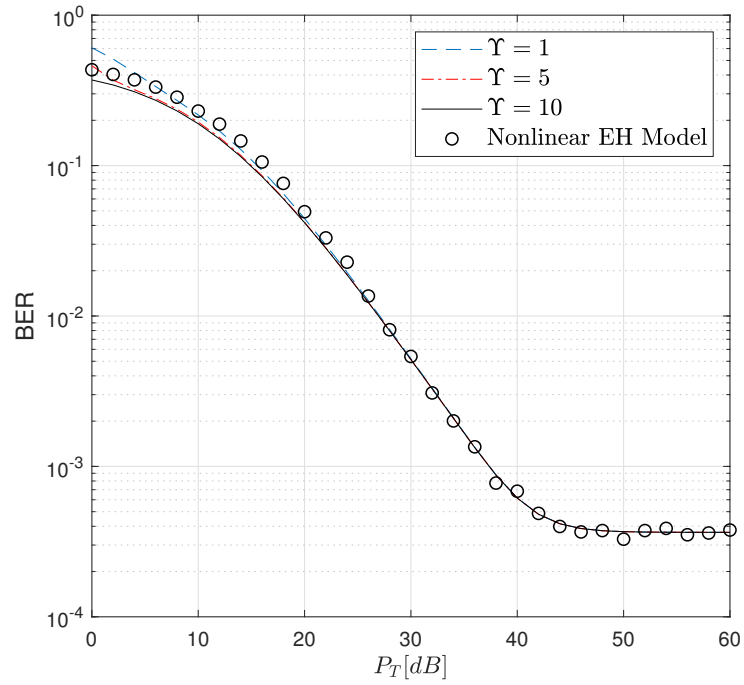


Figure 3.4 : BER performance of the WP DH AF relaying system versus P_T dB for $P_{th} = 35$ dB.

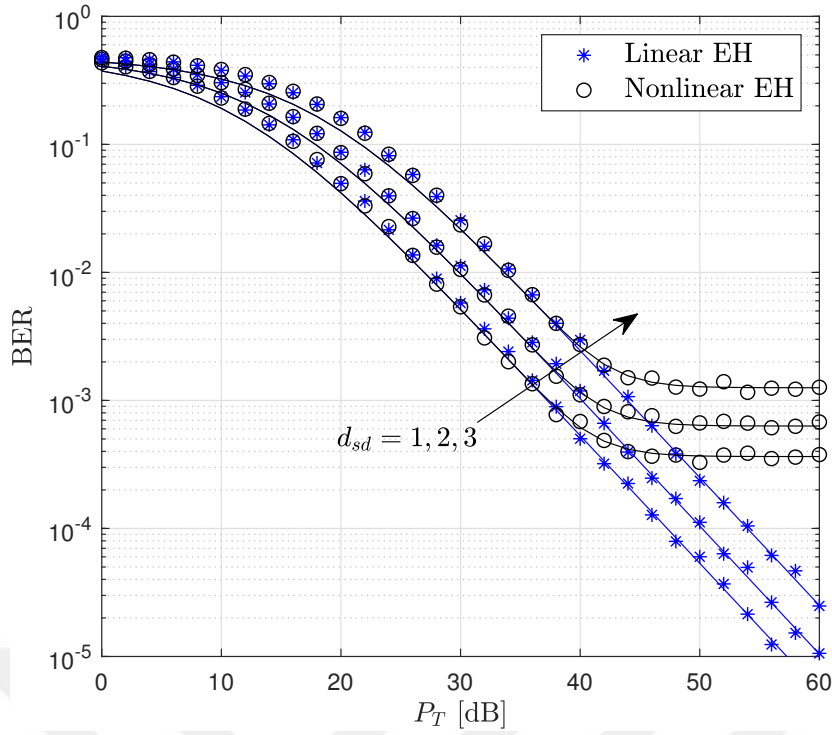


Figure 3.5 : BER performance of the WP DH AF relaying system versus P_T dB where $P_{th} = 35$ dB.

2 dB, and 6 dB SNR losses are obtained for the BER of 10^{-3} by increasing the value of d_{sd} from 1 to 2 and 3, respectively.

Figure 3.6 shows the BER performance versus d_h , where it is assumed that $d_{sd} = 2$. It is seen from Figure 3.6 that the results provided for $P_T = 30$ dB and $P_T = 40$ dB are similar for both linear and non-linear EH models, since the threshold power in the non-linear EH model is assumed as $P_{th} = 35$ dB, and the amount of harvested energy in both models is equal. However, for higher values of P_T , the linear EH model outperforms the non-linear EH where the amount of harvested power is saturated at $P_{th} = 35$ dB, which weakens the system performance compared to the linear EH model. Please note that for $P_T = 50$ dB and $P_T = 60$ dB, the results provided for the non-linear EH model are approximately the same, since for higher input power of PB, P_s and P_r are limited by $P_{th} = 35$ dB.

The throughput of the considered system versus P_T for $d_{sd} = 1$ is illustrated in Figure 3.7. It is shown that for a target rate value of $R = 1$ bit/sec/Hz, the linear and non-linear EH models provide approximately the same performance, while for both $R = 2$ and $R = 3$, the linear EH model has higher throughput compared to the

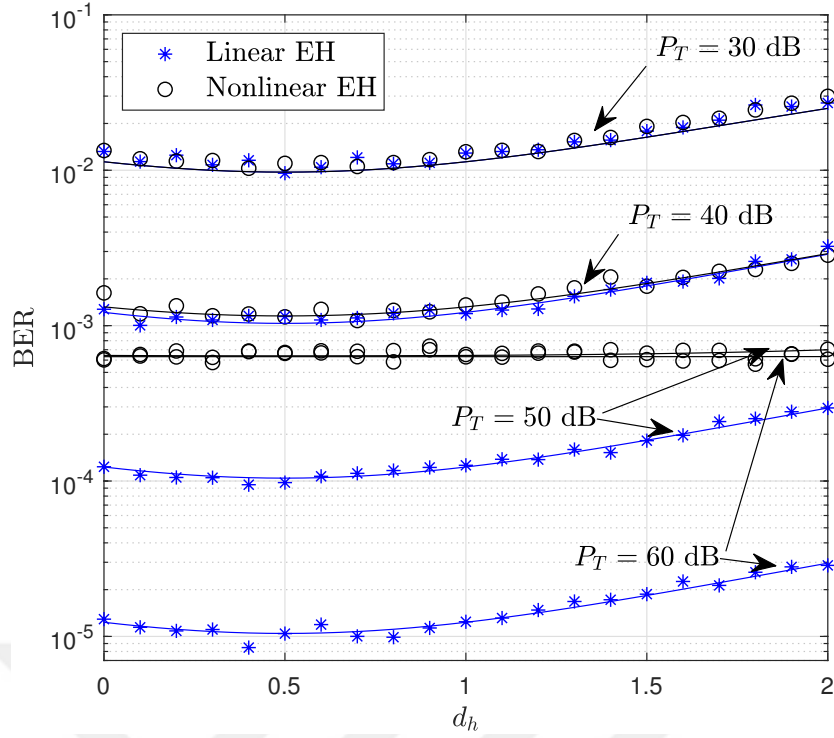


Figure 3.6 : BER performance of the WP DH AF relaying system versus d_h ($m_w = m_z = 2$, $P_T = \{30, 40, 50, 60\}$ dB).

non-linear EH model. This is based on the fact that the linear EH model causes a misrepresentation of the system performance compared to the considered EH model since the amount of the harvested energy at both S and R nodes is miscalculated. Hence, this provides a misunderstanding of the design of the EH systems.

Figure 3.8 depicts the throughput with respect to P_T for different values of d_{sd} . The target rate is fixed at $R = 1$. The results reveal that, for the distances $d_{sd} = 1, 2, 3$, the linear EH model provides performance approximately equivalent to that of the non-linear EH since the amount of harvested powers are approximately equal for both EH models. Moreover, in all cases, the throughput reaches its maximum value at $P_T = 42$ dB.

The throughput performance versus P_T for different values of P_{th} is shown in Figure 3.9. It is seen from Figure 3.9 that, by increasing P_{th} from 10 dB to 35 dB, the throughput increases and reaches its maximum value for $P_{th} = 35$ dB, which is approximately equivalent to the linear EH model throughput performance. In other words, a high level of threshold power in the non-linear EH model provides the same

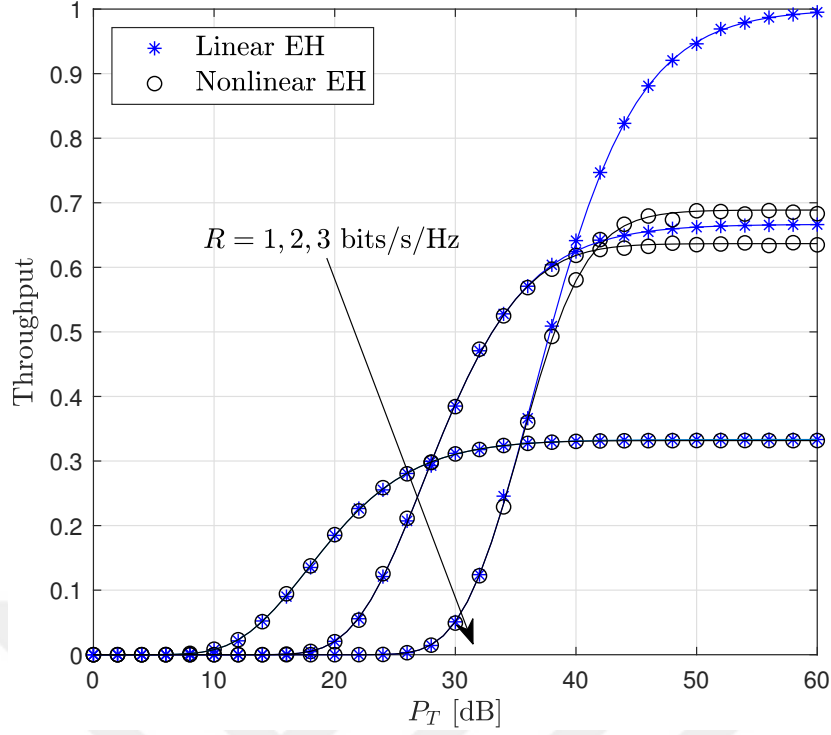


Figure 3.7 : Throughput performance of the WP DH AF relaying system versus P_T ($d_{sd} = 1$, $P_{th} = 35$ dB).

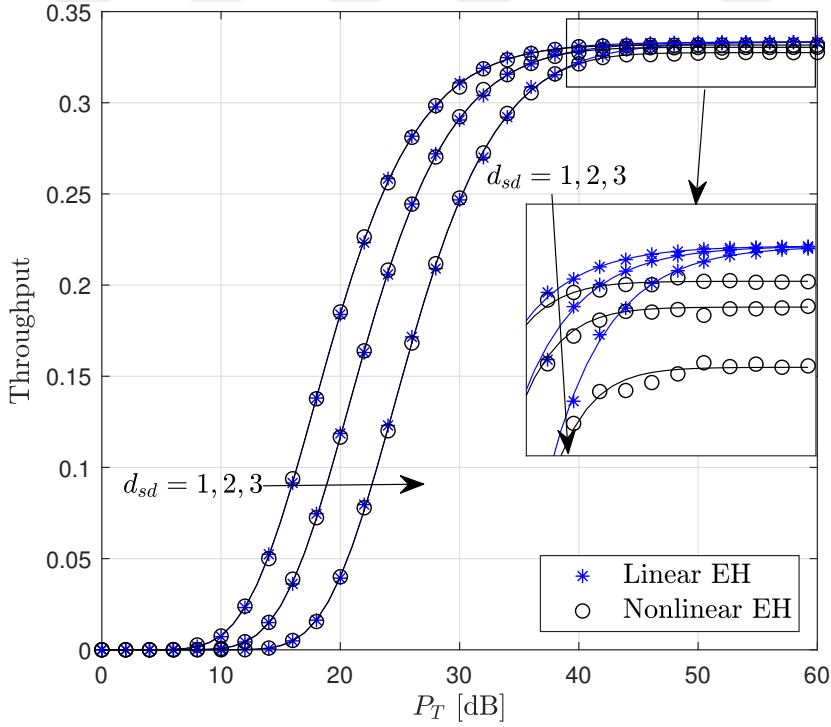


Figure 3.8 : Throughput performance of the WP DH AF relaying system versus P_T ($R = 1$ bits/s/Hz, $P_{th} = 35$ dB).

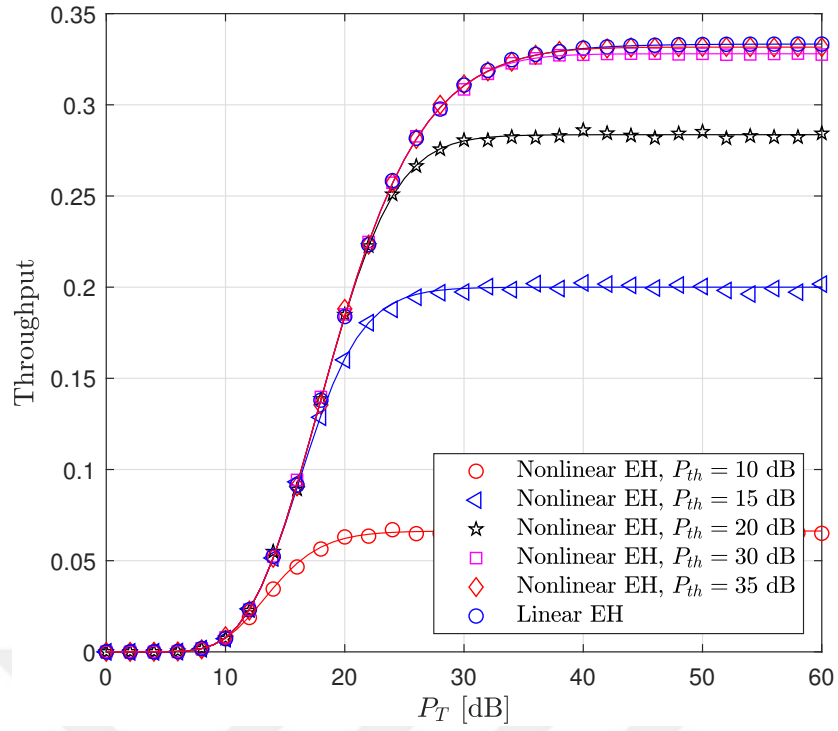


Figure 3.9 : Throughput performance of the WP DH AF relaying system versus P_T ($R = 1$ bits/s/Hz, $d_{sd} = 1$).

results as in the linear EH model since the amount of harvested power for both EH models is equal.

Figure 3.10 plots the source and relay harvested powers versus P_T . Please note that, from Appendix J, for an equal distance of S and R from PB, and for equal fading channel parameters, the harvested powers P_s and P_r are equal. The curves in Figure 3.10 are obtained using (J.1) and (J.2) for linear and non-linear EH models, respectively. As seen from Figure 3.10a,b, the distance (d_{sd}) and energy efficiency (η) have a significant effect on the amount of powers harvested at S and R. It is observed that, for high values of P_T , the average harvested power in the linear EH model is higher compared to the considered non-linear EH model, which is saturated to a predefined threshold value P_{th} . This high value is contrary to the amount of harvested energy in practical NL EH models, where it results in an overestimation of system performance.

Figure 3.11 shows the BER performance versus channel estimation error power σ_{err}^2 for different SNR values. Both linear and non-linear EH results are obtained considering $P_{th} = 35$ dB and $d_{sd} = 1$. Moreover, we modeled the imperfect CSI case as $\hat{g}_1 = g_1 + e$ and $\hat{g}_2 = g_2 + e$, where e denotes the the channel estimation error with distribution

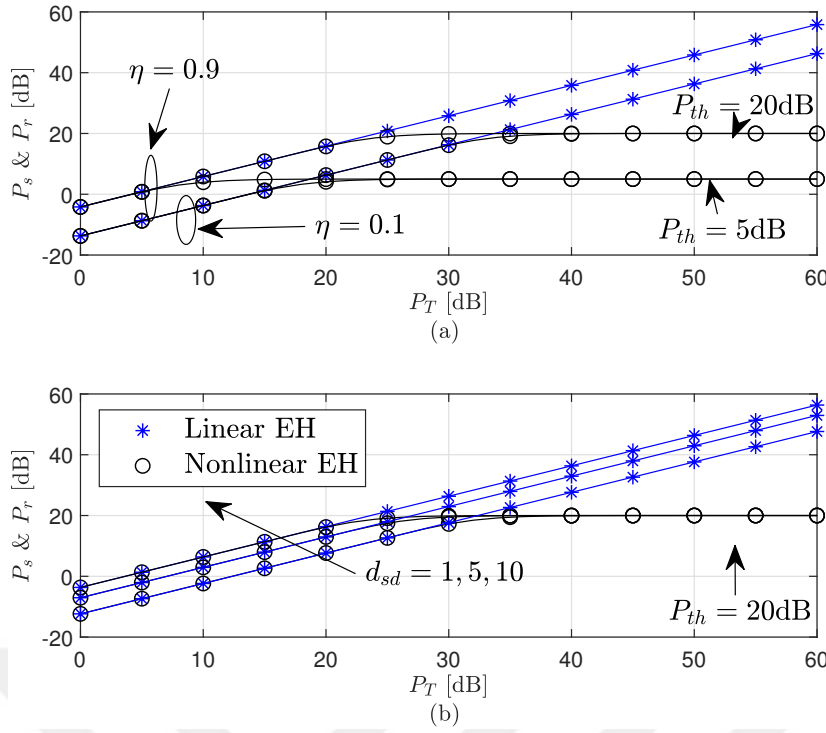


Figure 3.10 : P_s and P_r of the WP DH AF relaying system (a) $d_{sd} = 2$, (b) $\eta = \eta_s = \eta_r = 0.9$.

$\mathcal{CN} \sim (0, \sigma_{err}^2)$, σ_{err}^2 denoting the power of the channel estimation error. It is shown that system performances get worse when the power of channel estimation error increases for both linear and non-linear EH models.

3.7 Conclusions

In this chapter, a WP DH AF relaying system has been considered and a comprehensive performance analysis has been undertaken considering linear and non-linear EH models. In the studied system, S and R have been assumed to be battery-less, and that they harvest their power from a dedicated PB. For a comprehensive analysis, the results have been obtained considering the non-linear EH model together with the linear EH model which is mostly assumed in the literature. The BEP and throughput of the system for both linear and non-linear EH models have been analytically derived and compared with Monte Carlo simulation results, which verify our theoretical derivations. Moreover, results have been provided for different system parameters. It has been shown that the linear EH model misrepresents the system performance at high amounts of harvested energy and only provides reasonable results at low amounts of harvested energy. Hence, this causes an inaccurate understanding of the process

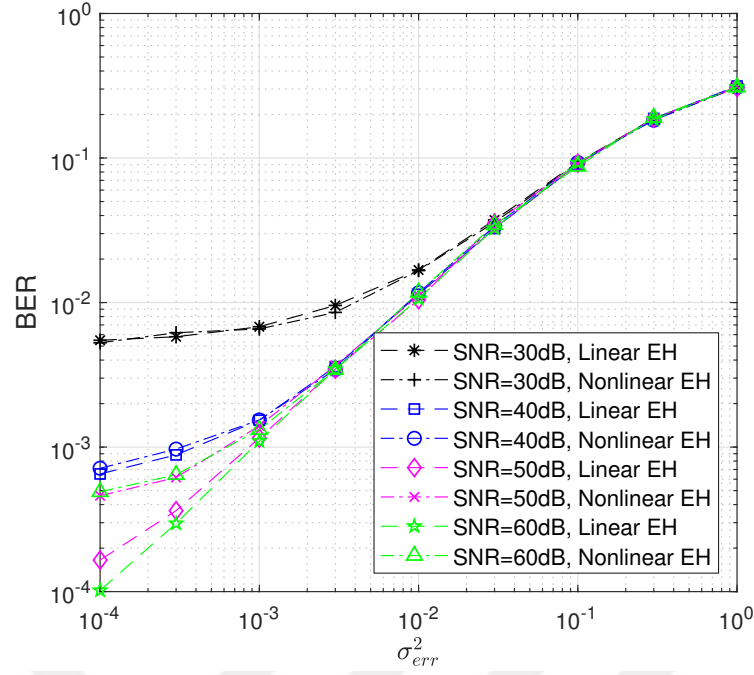


Figure 3.11 : BER performance of the WP DH AF relaying system versus P_T dB where $P_{th} = 35$ dB, $d_{sd} = 1$.

of EH system design. However, the studied non-linear EH model provides a realistic result for the system performance, both at low and high harvested energy levels, which provides a comprehensive insight into the EH system architecture. We considered the case of a single antenna in all nodes in this chapter. However, it is possible to have more antennas at the S, R, and D nodes and to analyze the improved system performance based on transmit antenna selection and maximum ratio combining. The analysis of these systems will be part of future work.



4. NEW PRACTICAL NONLINEAR ENERGY HARVESTING MODELS FOR WIRELESS-POWERED COMMUNICATIONS³

As a massive number of nodes are expected to be connected to each other in next-generation wireless networks, the power supplying of such multiple-antenna nodes is a challenge and requires new and sustainable solutions. Radio frequency energy harvesting (EH) is among the promising approaches to solve this issue. In this chapter, two new mathematically tractable nonlinear EH (NL-EH) models are proposed, for which the system performance is evaluated in terms of average harvested power, throughput, and bit error probability. The system consists of one multi-antenna power-constraint source harvesting its power from a dedicated power beacon and transmitting its signal to a destination equipped with multiple antennas. For a comprehensive analysis of the system, closed-form expressions are derived for Nakagami- m fading channels and the special case of Rayleigh channels. For comparison purposes, the performance derivations in Nakagami- m fading channels are provided for linear and piece-wise linear EH models given in the literature. Besides, the simulation results are obtained applying Monte-Carlo method for NL-EH models existing in the literature. The results provide a broadening view of EH systems and thoroughly compare the proposed NL-EH models with linear, piece-wise linear, and NL EH models available in the literature. Hence, these provide better insights towards the analyses and design of EH systems.

4.1 Introduction

In next-generation wireless communications, a huge number of nodes with multiple antennas will be connected to each other, and power requirements will be the critical constraint for sustainability. Wireless energy harvesting (EH) methods are considered among promising solutions to the energy shortage problem of these nodes. Considering

³This chapter is based on "Babaei, M., Durak-Ata, L. and Aygözü, Ü.(2022). New Practical Nonlinear Energy-Harvesting Models for Wireless-Powered Communications, Wireless Communications and Mobile Computing, vol. 2022, pp. 19, doi: 10.1155/2022/6534754."

the constraint energy resources and green communication, radio frequency (RF) EH provides a small step to solve the energy shortage because it replaces the usage of batteries and electronic power supplies at the sensor nodes. The EH node is equipped with an EH circuit which provides the ability to harvest the incoming signal energy. Wireless power transfer (WPT) and simultaneous wireless information and power transfer (SWIPT) are two EH methods considered in the design of battery-less nodes [14, 21, 81]. Based on the EH circuit feature, the power-constraint node can harvest the total incoming signal power or simultaneously harvest energy and process the information [14, 21]. Power splitting (PS) and time switching (TS) are examples of SWIPT method. In the WPT, dedicated power beacon (PB) or ambient power nodes are the sources that the power-constraint nodes harvest their power [81]. In the above-mentioned type of EH circuits and EH sources, the important factors are the amount of harvested power and the characteristic of EH circuit which directly affect the EH system performance. Linear, piece-wise linear, and nonlinear EH (NL-EH) models characterize the amount of harvested power [6–8, 10, 14, 82]. Most works in the literature consider the linear EH model which has a direct proportion between the input and output powers of the EH circuit [14, 83, 84]. On the other hand, the piece-wise linear EH model is considered in the literature [8, 82] which provides the same amount of energy as in the linear model for the small amount of input power. Different from the linear EH model, it saturates the output power to a fixed maximum value at high input powers.

With the measured data from the practical experimental tests given in [1–3, 85–95], it is figure out that the input and output powers relation of the EH circuit is nonlinear and the output power saturates to a maximum power even increasing the input power. Applying the curve fitting over the measurement data some NL-EH models are proposed in the literature [6, 7, 10]. This characteristic of the EH node is different from the linear EH model in which the harvested power is directly proportional to the input power and the harvested power increases with rising the input power of the EH circuit. Consequently, this results in an overestimation of the amount of harvested power and this yields an inaccurate understanding of EH system performance. The system performance in terms of outage probability and bit error probability (BEP) is investigated well for the linear EH models [14, 21] and the piece-wise linear

EH model [81] given in the literature. However, the system performance of the NL-EH models is mostly studied in terms of the amount of harvested energy and its statistical characteristics [6, 10, 96] since these NL-EH models are not mathematically tractable when investigating their bit error performances. Additionally, a small number of studies investigated the throughput [7]. This problem comes from the fact that for the NL-EH models given in the literature it is impossible to calculate the probability density function (PDF) or cumulative density function (CDF) of the received signal-to-noise ratio (SNR), which prevents the derivations of BEP or throughput.

The ergodic capacity and throughput of dual-hop (DH) amplify-and-forward (AF) relaying system obtained from outage analysis are investigated in [14]. The results are provided for PS, TS, and ideal EH modes where a linear EH model is applied at the power-constraint node. The results are obtained for different parameters and the authors show the discrepancy of the PS and TS EH modes in terms of the throughput from the ideal EH mode. The closed-form expressions for outage probability and BEP are derived in a SWIPT based DH AF and DH decode-and-forward (DF) relaying systems [21]. The channels are subject to Nakagami- m fading and the power-constraint relay node applies the linear EH model. The performance is provided for different system parameters and generally, PS EH mode outperforms the TS EH mode. System performance of the WPC DH AF relaying system is investigated in [97] where both S and R are power-constraint nodes and harvest their power from a dedicated PB. BER of the system is obtained for different parameters, and theoretical derivations are derived considering the linear EH model.

A DH AF relaying system with a power-constraint relay is considered in [8] where the relay applies TS mode to harvest its power based on the piece-wise linear EH model. The outage capacity of the studied system is derived considering Nakagami- m fading channels. A multi-relay system where source (S) and destination (D) communicate with each other with the help of power-constraint relay nodes is investigated under the Rayleigh fading channel assumption [98]. Outage probability and throughput of the system are derived for the piece-wise linear EH model and imperfect channel state information (CSI). In the studied system, both transmit antenna and relay selections are considered for the transmission protocol. In [81], an overlay cognitive radio with

a power-constraint secondary transmitter (ST) is investigated. The authors assumed that ST harvests its power from the primary transmitter and applies the piece-wise linear EH model. The closed-form expressions for the outage probabilities and BEPs are derived considering both linear and piece-wise linear EH models which provide a good insight for comparison purposes.

The throughput of DH AF relaying system considering Rayleigh fading is studied in [82]. TS and PS EH mode with a segmented piece-wise linear EH model is applied in the battery-less relay node. The authors show that the PS mode outperforms the TS mode. SWIPT based cooperative non-orthogonal multiple access (NOMA) system performance is studied in [99] where the near user is power-constraint and applies a segmented piece-wise linear EH model. The theoretical derivations of outage probability and system throughput are provided for the Rayleigh fading channel.

In [6], a NL-EH model based on logistic (sigmoidal) function is proposed. Assuming a non-convex optimization and simplifying it to a convex optimizing problem, the total harvested power is maximized and the average harvested power is calculated. The rate-energy trade-off region considering both linear and NL-EH models is obtained in [100]. The harvested power versus the number of receive antennas is investigated. In [101], average harvested power of the point-to-point MIMO WPT system is calculated based on different system parameters. The authors study the maximization of harvested power jointly by optimizing the PS ratio factor and the transmit energy of PB. A MISO NOMA SWIPT system is investigated in [102] where the authors study the physical layer security by adopting the NL-EH model from [6]. The authors simplify the non-convex optimization problem to a convex one.

The quadratic NL-EH model is proposed and studied in a multi-user orthogonal frequency-division multiple access (OFDMA) system [10]. Results are given compared to the linear EH model. Moreover, it is assumed that the parameters of proposed function in [10] take values of $\alpha_1 < 0$ and $\alpha_2 > 0$, and $\alpha_3 < 0$ which is shown in Table 4.1. However, a high input power results in a negative value for harvested power. Hence, this proposed model is only valid for small input powers. A heuristic NL-EH model is proposed in [7] applying curve fitting. The throughput of AF cooperative relaying system using PS and TS modes is investigated where the results are numerically obtained since the proposed model is not mathematically tractable.

The NL-EH models given above are based on the curve fitting of experimental measured data. The harvested powers given in [6, 7, 10] include the polynomial, exponential or cubic functions of the input power. Consequently, all these NL-EH models given in the literature are not mathematically tractable and the harvested power does not provided as PDF and CDF to further calculate the EH system performance. These are the main motivations behind our article.

In this chapter, we propose two NL-EH models which are mathematically tractable in terms of the PDF, CDF of harvested power, and the system performance at the expense of a small gap in the accuracy between measured data points and those from our models. Apart from the NL-EH models given in the literature, our models depend directly on the system input power rather than the polynomial, exponential or cubic functions of it. The considered system consists of a PB from where a power-constraint S harvests energy which is then used to transmit data to a D. The channels are assumed subject to Nakagami- m and Rayleigh fading. The former is assumed when the channels between nodes are stronger and the distances are short. In the later, the link between nodes is subjected to the deleterious effect of the environment in the lack of a line-of-sight component.

For both models, PDF and CDF expressions for the harvested power are obtained based on the curve fitting of the measurements data from [1, 2]. Moreover, the average harvested powers are calculated using the PDF of harvested power. Furthermore, the outage probability, throughput, and BEP of the considered system are derived in closed-forms for the proposed two NL-EH models. Note that the NL-EH models considered in the literature do not provide closed-form expressions for the system performance since they are not mathematically tractable [6, 7, 10]. In the papers considering these models, only the amount of harvested power at the EH receiver is studied without investigating the system BEP or throughput performance.

Numerical results provide a comprehensive analysis for all the considered EH models in terms of system performance. The results show that the distance and number of antennas have crucial effects on the system performance which gives an insight into maximizing the system performance. The main contributions of the chapter are as follows:

- Two mathematically tractable NL-EH models based on the curve fitting of measured data are proposed.
- PDF and CDF of the harvested power for the proposed NL-EH models are provided.
- Closed-form expressions for the average harvested power, throughput and BEP are derived and the results are supported by Monte-Carlo simulations.
- Theoretical derivations of the system performance are provided for both Nakagami- m and Rayleigh fading.
- Results are compared with existing linear, piece-wise linear, and NL-EH models from the literature.

The chapter is organized as follows. Section 4.2 describes the considered system model. The proposed NL-EH models are presented in Section 4.3. Expressions for the average harvested powers are obtained in Section 4.4. Section 4.5 deals with the derivation of throughput expressions. Bit error rate (BER) analyses are performed in Section 4.6. Theoretical and simulation results are presented in Section 4.7. Section 4.8 concludes the chapter.

4.2 System Model

The considered half-duplex RF EH system where a power-constraint S harvests its power from a PB is given in Figure 4.1. S and D are equipped with N_s and N_d receive antennas, respectively, while PB is equipped with one antenna. S transmits its signal from one antenna to D using the harvested power. d_{pbs} and d_{sd} denote the distances of links PB $\rightarrow$ S and S $\rightarrow$ D, respectively. Additionally, the path-loss is represented as $L_i = 1/(1 + d_i^\nu)$ where $i \in \{pbs, sd\}$, and ν is the path-loss coefficient. x_s is the S's transmitted signal with $\mathbb{E}\{|x_s|^2\} = 1$. $\mathbf{h} = [h_1, h_2, \dots, h_{N_s}]$ and $\mathbf{g} = [g_1, g_2, \dots, g_{N_d}]$ represents the channel gain vectors of links PB $\rightarrow$ S and S $\rightarrow$ D, respectively. Furthermore, $\mathbf{n} = [n_1, n_2, \dots, n_{N_d}]$ is the additive white Gaussian noise (AWGN) vector with entries as i.i.d complex Gaussian random variable $n_i \sim \mathcal{CN}(0, N_0)$ for $i \in \{1, 2, \dots, N_d\}$. Transmission of the S signal takes T seconds where $T/2$ seconds are allocated for harvesting energy and the remainder for IP. It is assumed that the CSI is perfectly known at node D and channels are assumed to be flat

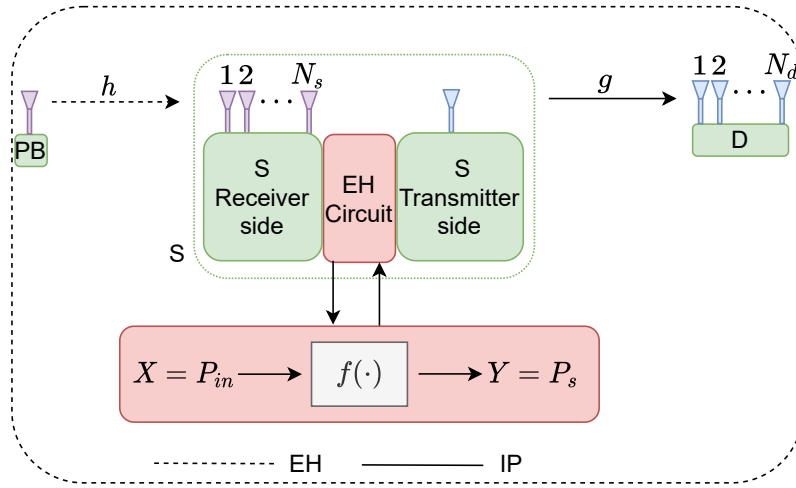


Figure 4.1 : Considered half-duplex wireless-powered system model where $\mathbf{h}$, and $\mathbf{g}$ represent channel gain vectors. P_{in} , and P_s are the input and output power of the EH circuit, respectively. $f(\cdot)$ is the function converting the input power of the EH circuit to the output power.

fading and remain constant during each transmission interval and vary independently from an interval to another. Line-of-sight (LoS) and distance are two determinant factors considered in the RF EH communication systems. Since the path-loss causes a substantial decrease in the amount of harvested power, the EH communication is possible for short distances and in the presence of a LoS component. Generally, Rician fading is a proper channel model with its LoS component. However, mathematical derivations of the system performance considering the Rician PDF and CDF are not tractable. Meanwhile, Nakagami- m fading is a good approximation of Rician model [33, 103]. Hence, it is assumed that both links $\text{PB} \rightarrow \text{S}$ and $\text{S} \rightarrow \text{D}$ are subject to Nakagami- m fading with parameters m_h and m_g , respectively. Here, we consider that the channel power is identical for each link. In other words, $\Omega_h = E\{|h_i|^2\}$ and $\Omega_g = E\{|g_j|^2\}$ for $i \in \{1, 2, \dots, N_s\}$, and $j \in \{1, 2, \dots, N_d\}$, respectively. In the first time slot, the input power of the EH circuit provided in Figure 4.1 is calculated as

$$P_{in} = P_T L_{pbs} \left(\sum_{i=1}^{N_s} |h_i|^2 \right) \quad (4.1)$$

where P_T is the energy-bearing signal transmitted from PB. In the second time slot, the received signal at D is given as

$$\mathbf{y} = \sqrt{P_s L_{sd}} \mathbf{g} x_s + \mathbf{n} \quad (4.2)$$

where $\mathbf{y} = [y_1, y_2, \dots, y_{N_d}]$, $\mathbf{g} = [g_1, g_2, \dots, g_{N_d}]$, and P_s is the transmit power of S. Considering (4.1), we have

$$P_s = f(P_{in}) \quad (4.3)$$

where $f(\cdot)$ is given in Figure 4.1. The received SNR after applying maximum ratio combining (MRC) at node D is given using (4.2) as

$$\gamma = YZ \quad (4.4)$$

where $Z = L_{sd} \sum_{i=1}^{N_d} |g_i|^2 / N_0$. For simplicity, we define $X = P_{in}$, and $Y = P_s$ in the sequel. The CDF, and PDF of r.v.s Z and X are given as [33]

$$F_Z(z) = 1 - \sum_{i=0}^{m_z-1} \frac{(z\phi_z)^i}{i!} \exp(-z\phi_z), \quad (4.5)$$

and

$$f_X(x) = Dx^{m_x-1} \exp(-x\phi_x), \quad (4.6)$$

respectively, where $\phi_z = m_z/\bar{z}$, $\bar{z} = L_{sd}\Omega_z/N_0$, $\Omega_z = N_d\Omega_g$, $m_z = N_d m_g$, $D = \phi_x^{m_x}/\Gamma(m_x)$, $\phi_x = m_x/\bar{x}$, $\bar{x} = P_T L_{pbs}\Omega_x$, $\Omega_x = N_s\Omega_h$, and $m_x = N_s m_h$. Here, we assume that the channel parameters of Nakagami- m fading follow a positive integer in (4.5).

4.3 Linear and NL-EH Models

In the literature, different linear and NL-EH models are provided for the performance evaluation of EH communication systems [6–8, 10, 14, 82]. These EH models are listed in Table 4.1. For the linear EH model, the harvested power is linearly dependent on the input power of the EH circuit with a constant ratio parameter, namely, the conversion efficiency (η). However, based on the data measurements from designed practical EH circuits, linear EH models overestimate the amount of the harvested energy especially for high input powers which consequently results in a misunderstanding of the system performance. Moreover, the EH circuits have nonlinear properties which can be modeled by a nonlinear function of the amount of harvested power with respect to the input power. Furthermore, the piece-wise linear EH model is provided in the literature where the harvested amount of energy is saturated to a threshold power as P_{th} while acts the same as the linear EH model for the small input powers as seen

Table 4.1 : Existing NL-EH and linear EH models.

EH models	Harvested Power P_s
Linear [14]	$Y = \eta X$
Piece-wise linear [8]	$Y = \begin{cases} \eta X, & X < P_{th} \\ \eta P_{th}, & X > P_{th} \end{cases}$
i -Segment	$Y = \begin{cases} 0, & 0 < X < P_{th_1} \\ a_1 X + b_1, & P_{th_1} < X < P_{th_2} \\ \cdots, & P_{th_{i-1}} < X < P_{th_i} \\ P_{th_i}, & X > P_{th_i} \end{cases}$
Piece-wise linear [82]	
NL-EH [10]	$Y = \alpha_1 X^2 + \alpha_2 X + \alpha_3$
NL-EH [6]	$Y = \frac{\frac{M}{1 + e^{-a(X-b)}} - \frac{M}{1 + e^{ab}}}{1 - \frac{1}{1 + e^{ab}}}$
NL-EH [7]	$Y = \frac{p_2 X^3 + p_1 X^2 + p_0 X}{q_3 X^3 + q_2 X^2 + q_1 X + 1}$

from Table 4.1. In addition to these EH models, NL-EH models are also proposed applying curve fitting from the measurement data experiments from different EH circuits (Table 4.1). Please note that the theoretical derivations of system performance are not mathematically tractable considering these known NL-EH models. However, the linear and piece-wise linear EH models are mathematically tractable and provide a general and simple analysis of the system. On the other hand, the proposed NL-EH models in the sequel, enable mathematically tractable theoretical derivations of system performance. This provides an advantage on the design and evaluation of EH systems.

4.3.1 Proposed NL-EH models

In this subsection, two EH models are proposed with their relations between input and output powers. The first proposed NL-EH model (NL-EH-1) is given as

$$Y = f(X) = \frac{MeX - bc}{X + Mb/e}. \quad (4.7)$$

Here, M denotes the maximum saturation power of the EH circuit, b , c and e are constants to be determined by curve fitting. Applying (4.7), the CDF of the harvested

power of NL-EH-1 related to the input power is calculated as

$$F_Y(y) = Pr\left(\frac{aX - bc}{X + d} < y\right) = Pr(X(a - y) < yd + bc)$$

$$= \begin{cases} F_X\left(\frac{yd + bc}{a - y}\right), & -bc/d < y < a \\ 1, & y > a \end{cases} \quad (4.8)$$

where $a = Me$, $d = Mb/e$, and

$$F_X(x) = 1 - \sum_{i=0}^{m_x-1} \frac{(x\phi_x)^i}{i!} \exp(-x\phi_x). \quad (4.9)$$

Moreover, the PDF of NL-EH-1 is calculated using (4.8) as

$$f_Y(y) = \begin{cases} \frac{da + bc}{(a - y)^2} f_X\left(\frac{yd + bc}{a - y}\right), & -bc/d < y < a \\ 0, & y > a \end{cases} \quad (4.10)$$

where $f_X(x)$ is given in (4.6).

The second proposed NL-EH model (NL-EH-2) is given as

$$Y = f(X) = M \frac{cX}{X + b}. \quad (4.11)$$

Similar to NL-EH-1, the CDF of harvested power is obtained as

$$F_Y(y) = Pr\left(\frac{aX}{X + b} < y\right) = Pr(X(a - y) < yb)$$

$$= \begin{cases} F_X(yb/(a - y)), & 0 < y < a \\ 1, & y > a \end{cases} \quad (4.12)$$

where $a = Mc$. Taking the derivation of (4.12), the PDF of NL-EH-2 is calculated as

$$f_Y(y) = \begin{cases} \frac{ba}{(a - y)^2} f_X(yb/(a - y)), & 0 < y < a \\ 0, & y > a. \end{cases} \quad (4.13)$$

Figures 4.2 and 4.3 describe the curve fitting of the measured experimental data compared to the linear EH and the NL-EH models existing in the literature and the two proposed models in (4.7) and (4.11). The results of the curve fitting are obtained based on the measured data from [1] and [2]. The measured data are denoted by symbols while the fitted data represented by lines and dashed lines. It is seen from both Figures 4.2 and 4.3 that both the proposed NL-EH models fit better to data points than the linear and piece-wise linear models [8, 14]. However, as stated, the challenging

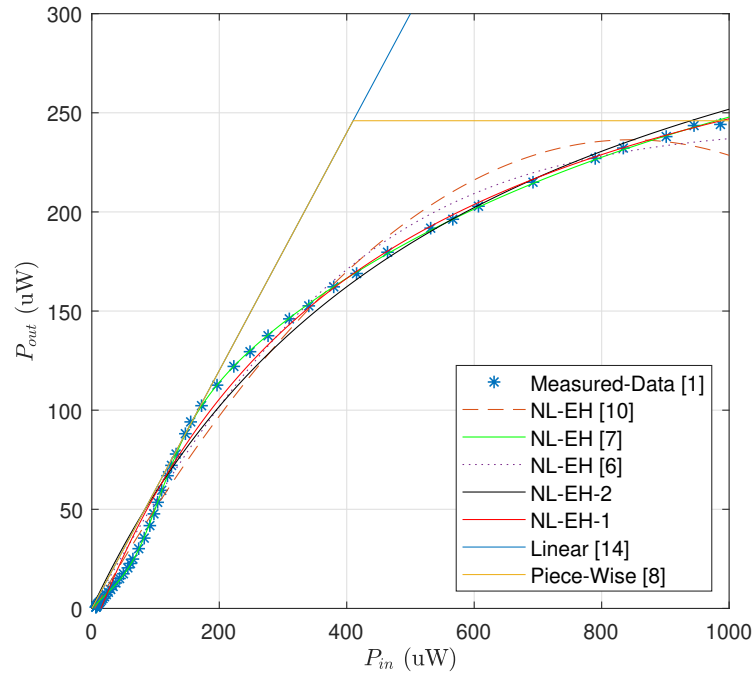


Figure 4.2 : Harvested power versus input power comparison of the proposed and existing EH models with experimental results from [1]

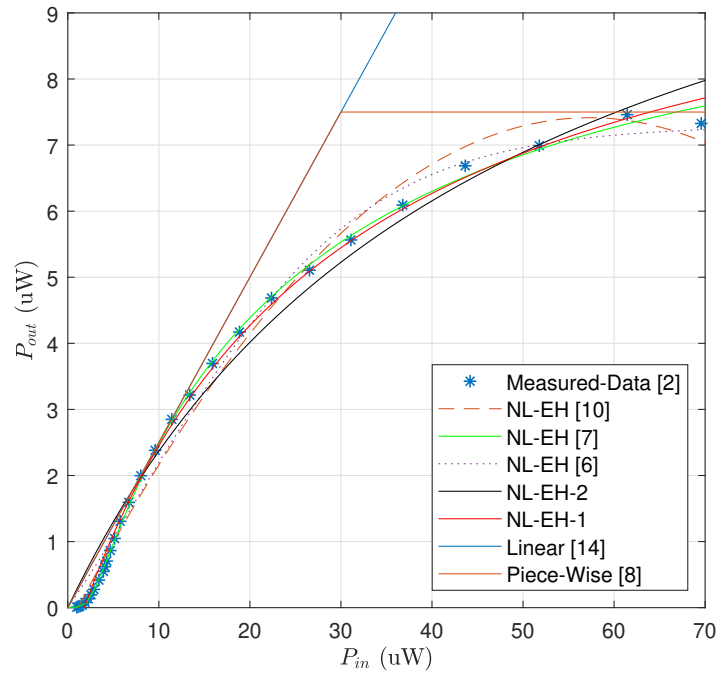


Figure 4.3 : Harvested power versus input power comparison of the proposed and existing EH models with experimental results from [2]

Table 4.2 : Goodness comparison of curve Fitting for different NL-EH models with data measurements taken from [1] and [2].

EH Models	$\mathcal{R}^2$, [1]	$\mathcal{R}^2$, [2]
NL-EH-1 (4.7)	0.9997	0.99994
NL-EH-2 (4.11)	0.99013	0.98066
NL-EH [6]	0.99277	0.99126
NL-EH [10]	0.98979	0.99328
NL-EH [7]	0.9999	0.99901

Table 4.3 : Fitting parameters of the proposed NL-EH-1 (4.7) in different energy harvesting circuits.

References	M	e	b	c
[1]	244.14	1.46807	2.58518	2003.91475
[3]	24.64643	1.30245	0.50571	299.49425
[85]	9.49	1.26804	4.41531	6.40408
[86]	994.72	1.54374	2.09621	0.01172
[2]	7.33	1.48839	5.3324	3.70781
[87]	87.32	5.12769	51.63326	$2.14287e - 4$
[88]	71.42	1.18517	2.90684	$6.45064e - 4$
[89]	407.73	2.13149	7.91512	0.00272
[90]	503.95	2.11327	6.71826	0.00122
[91]	156.83	15.38538	442.76435	$1.25506e - 5$
[92]	225.06	1.64925	3.53621	$6.04258e - 4$
[93]	594.75	3.8096	56.63872	0.00194
[94]	924.49	3.10906	13.73995	0.00653
[95]	345.393	1.6489	2.86208	0.03637

issue of the existing NL-EH models are the difficulties in mathematical derivations of CDF and PDF of the harvested power, and theoretical derivations of throughput as well as BER system performances. In addition, it is observed that apart from the NL-EH model in [10], the other NL-EH models provide similar curve fitting functions. Moreover, NL-EH model in [7] provides the best fit compared to all other NL-EH models. However, despite these similarities, these models given in the literature are not mathematically tractable which prevent theoretical investigation of the EH system performance. Meanwhile, the proposed NL-EH models which have an acceptable goodness in fitting, are mathematically tractable.

Moreover, the goodness of the fits considering the NL-EH models are provided in Table 4.2 for [1] and [2], respectively. $\mathcal{R}^2$ (R-squared) is the parameter that is statistically calculated from curve fitting functions and data measurements [104]. The strength of the fit of a model is most commonly evaluated using $\mathcal{R}^2$ which always takes values

Table 4.4 : Fitting parameters of the proposed NL-EH-2 (4.11) in different energy harvesting circuits.

References	M	c	b
[1]	244.14	1.63202	582.15452
[3]	24.64643	1.51448	26.8605
[85]	9.49	1.39447	0.04939
[86]	994.72	1.58195	1.46469
[2]	7.33	1.80594	45.66596
[87]	87.32	1.18247e14	2.96048e13
[88]	71.42	1.37742	0.32895
[89]	407.73	2.54365	2.12182
[90]	503.95	2.2353	1.77768
[91]	156.83	33.96747	10.38453
[92]	225.06	1.77004	0.56606
[93]	594.75	5.5573	14.38647
[94]	924.49	4.0876	6.04377
[95]	345.393	1.66525	6.17207

between 0 and 1. This statistical parameter measures how well the data is close to being fit. The closer the value to 1, the fit is better. Considering both Figures 4.2 and 4.3 and Table 4.2, the proposed NL-EH model functions provide similar results as their counterparts given in the literature. However, the important factor in choosing the NL-EH model is having mathematical tractability while having a well-fitting to the data measurements. In other words, a trade-off of between accuracy and mathematical tractability is considered. Furthermore, the coefficient parameters of NL-EH-1, and NL-EH-2 considered in (4.7), and (4.11) are obtained with curve fitting and provided in Tables. 4.3 and 4.4, respectively. As it is provided in both Tables. 4.3 and 4.4, the parameters of these models are obtained for measurement results of [1–3, 85–95].

Figure 4.4 and Figure 4.5 show the histograms of the PDFs of NL-EH-1 and NL-EH-2 given by (4.10), and (4.13) for (4.6), respectively. Moreover, the histograms of both PDFs are provided for different channel powers and parameters where it is assumed that $N_s = 1$, and $\Omega_1 = \Omega$. It is seen from both Figure 4.4 and Figure 4.5 that the histograms overlay the theoretical PDFs for the intended distributions. This verifies the accuracy of the derivations of both PDFs for NL-EH models. Please note that for all cases, the area of shaded regions are always equals to one which verifies the PDF calculations given in (4.10) and (4.13). Also, note that both histograms are bounded to the given boundaries defined in (4.10) and (4.13), respectively.

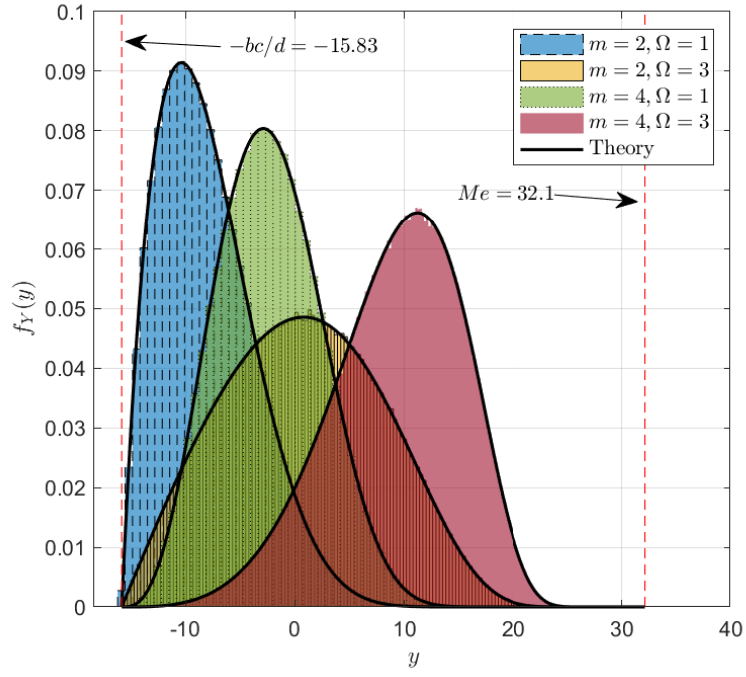


Figure 4.4 : Histogram and theoretical PDF of (a) NL-EH-1 where data measurements are taken from [3], and $N_s = 1$, $\Omega = \Omega_1$.

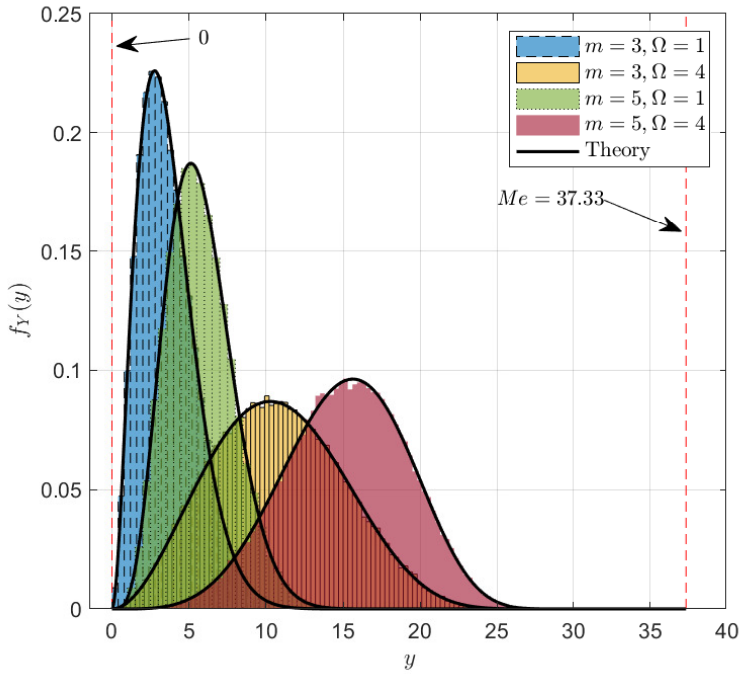


Figure 4.5 : Histogram and theoretical PDF of (b) NL-EH-2 where data measurements are taken from [3], and $N_s = 1$, $\Omega = \Omega_1$.

4.4 Average Harvested Power of the Proposed NL-EH Models

In this section, average harvested power of the two proposed NL-EH models are investigated. Additionally, for comparison purposes, average harvested power of the linear and piece-wise linear EH models are calculated.

4.4.1 Average harvested power of NL-EH-1 model

The average harvested power of NL-EH-1 model given in (4.7) is calculated as

$$\bar{Y}_1 = \int_0^\infty \frac{aX - bc}{X + d} f_X(x) dx = N_1 - N_2 \quad (4.14)$$

where

$$N_1 = \int_0^\infty \frac{aX}{X + d} f_X(x) dx \quad (4.15)$$

and

$$N_2 = bc \int_0^\infty \frac{1}{X + d} f_X(x) dx. \quad (4.16)$$

Here, $f_X(x)$ is given in (4.6). Substituting (4.6) in (4.15), (4.16) and applying [32, 3.383-10], we have

$$N_1 = Dad^{m_x} \exp(d\phi_x) \Gamma(m_x + 1) \Gamma(-m_x, d\phi_x) \quad (4.17)$$

and

$$N_2 = Dbcd^{m_x-1} \exp(d\phi_x) \Gamma(m_x) \Gamma(1 - m_x, d\phi_x). \quad (4.18)$$

where $\Gamma(\cdot, \cdot)$ is the upper incomplete Gamma function [32, 8.350-2]. Finally, the average harvested power for NL-EH-1 given in (4.7) is obtained by substituting (4.17) and (4.18) in (4.14).

4.4.2 Average harvested power of NL-EH-2 model

The average harvested power of NL-EH-2 model given in (4.11) is calculated as

$$\bar{Y}_2 = \int_0^\infty \frac{aX}{X + b} f_X(x) dx \quad (4.19)$$

where $f_X(x)$ is given in (4.6). Applying [32, 3.383-10] the average harvested power is obtained as

$$\bar{Y}_2 = Dab^{m_x} \exp(b\phi_x) \Gamma(m_x + 1) \Gamma(-m_x, b\phi_x). \quad (4.20)$$

4.4.3 Average harvested power of linear and piece-wise linear EH models

The average harvested powers of linear and piece-wise linear EH models given in Table 4.1 are written as

$$\bar{Y}_l = \int_0^\infty \eta x f_X(x) dx, \quad (4.21)$$

and

$$\bar{Y}_{pw} = \underbrace{\int_0^{P_{th}} \eta x f_X(x) dx}_{U_1} + P_M \underbrace{\int_{P_{th}}^\infty f_X(x) dx}_{U_2}, \quad (4.22)$$

respectively, where $P_M = \eta P_{th}$. Substituting (4.6) in (4.21), simplifying and applying [32, 3.326-2], the average harvested power of linear EH model for the considered system is obtained as

$$\bar{Y}_l = \eta D \Gamma(m_x + 1) / \phi_x^{m_x+1}. \quad (4.23)$$

Moreover, for the piece-wise linear EH model in (4.22), by substituting (4.6) we have

$$U_1 = \eta D \gamma(m_x + 1, \phi_x P_{th}) / \phi_x^{m_x+1} \quad (4.24)$$

and

$$U_2 = 1 - F_X(P_{th}) \quad (4.25)$$

where $F_X(\cdot)$ is given in (4.9). Here, (4.24) is calculated by [32, 3.351-1].

Based on the obtained expressions given in eqs. (4.14), (4.19), (4.21), and (4.22), the average harvested power depends on the PDF of the input power as well as the channel path loss parameter, number of harvesting antennas (N_s), and the transmit power of PB, as seen from (4.6).

4.5 Outage and Throughput Analyses

In this section, closed-form expressions are derived for the outage probability and throughput of the overall system $PB \rightarrow S \rightarrow D$ considering the both proposed NL-EH models in Nakagami- m fading channels. The throughput of the considered system is calculated as

$$\tau = (1 - F_\gamma(\gamma_{th}))R/2 \quad (4.26)$$

where $\gamma_{th} = 2^{2R} - 1$ and R bits/sec/Hz is the target rate. Moreover, $F_\gamma(\cdot)$ is the CDF of received SNR given in (4.4).

4.5.1 Outage probability and throughput of NL-EH-1 model

Considering (4.4) and (4.7), the conditional CDF of the received SNR is given as

$$F_\gamma(\gamma) = \begin{cases} F_Z\left(\gamma \frac{X+d}{aX-bc} | X\right), & bc/a < x < \infty \\ 1, & 0 < x < bc/a \end{cases} \quad (4.27)$$

where $F_Z(z)$ is given in (4.5). Averaging over X using (4.6), we have

$$F_\gamma(\gamma) = \int_0^{bc/a} f_X(x) dx + \int_{bc/a}^{\infty} F_Z\left(\gamma \frac{X+d}{aX-bc} | X\right) f_X(x) dx. \quad (4.28)$$

(4.28) is rewritten by substituting (4.5) and (4.6) as

$$F_\gamma(\gamma) = 1 - \lambda(i, \gamma) \int_{bc/a}^{\infty} \left(\frac{x+d}{ax-bc}\right)^i x^{m_x-1} \exp(-x\phi_x) \\ \times \exp\left(-\phi_z \gamma \frac{x+d}{ax-bc}\right) dx \quad (4.29)$$

where $\lambda(i, \gamma) = D \sum_{i=0}^{m_z-1} (\gamma \phi_z)^i / i!$. By defining $ax - bc = t$ and after some mathematical simplification, (4.29) is rewritten as

$$F_\gamma(\gamma) = 1 - \lambda(i, \gamma) \exp\left(-\phi_z \frac{\gamma}{a}\right) \exp\left(-\phi_x \frac{bc}{a}\right) \left(\frac{1}{a}\right)^{m_x+i} \\ \times \int_0^{\infty} t^{-i} (t+j)^i (t+bc)^{m_x-1} \exp\left(-\phi_x \frac{t}{a} - \phi_z \frac{\gamma J}{at}\right) dt \quad (4.30)$$

where $J = da + bc$. Using [32, 1.111], we have

$$\begin{aligned}
F_\gamma(\gamma) &= 1 - D \sum_{i=0}^{m_z-1} \frac{(\gamma\phi_z)^i}{i!} \exp\left(-\phi_z \frac{\gamma}{a}\right) \exp\left(-\phi_x \frac{bc}{a}\right) \\
&\quad \left(\frac{1}{a}\right)^{m_x+i} \sum_{k=0}^i \binom{i}{k} J^{i-k} \sum_{l=0}^{m_x-1} \binom{m_x-1}{l} (bc)^{m_x-1-l} \\
&\quad \times \int_0^\infty t^{k+l-i} \exp\left(-\phi_x \frac{t}{a} - \phi_z \frac{\gamma J}{at}\right) dt.
\end{aligned} \tag{4.31}$$

Applying [32, 3.471-9], the CDF of received SNR at D is calculated as

$$\begin{aligned}
F_\gamma(\gamma) &= 1 - \varpi(i, k, l) \gamma^{(u+i)/2} \exp\left(-\phi_z \frac{\gamma}{a}\right) \\
&\quad \times K_{u-i}\left(2\sqrt{\frac{\phi_x \phi_z J}{a^2}} \gamma\right)
\end{aligned} \tag{4.32}$$

where $u = k + l + 1$ and

$$\begin{aligned}
\varpi(i, k, l) &= 2D \left(\frac{1}{a}\right)^{m_x} \exp\left(-\phi_x \frac{bc}{a}\right) \sum_{i=0}^{m_z-1} \frac{(\phi_z/a)^i}{i!} \\
&\quad \times \sum_{k=0}^i \binom{i}{k} J^{i-k} \sum_{l=0}^{m_x-1} \binom{m_x-1}{l} (bc)^{m_x-1-l} \\
&\quad \times \left(\frac{\phi_z J}{\phi_x}\right)^{(k+l-i+1)/2},
\end{aligned} \tag{4.33}$$

and $K(\cdot)$ is the modified Bessel function of the second kind [32, 8.407-2]. The throughput of the considered system for NL-EH-1 model given in (4.7) is calculated by substituting $F_\gamma(\gamma)$ from (4.32) in (4.26).

Special case of Rayleigh fading for NL-EH-1 model For the case where both links from PB→S and S→D are exposed to Rayleigh fading and both S and D are equipped with one antenna, the CDF and PDF of (4.5) and (4.6) are simplified as

$$F_Z(z) = 1 - \exp(-z/\bar{z}), \tag{4.34}$$

and

$$f_X(x) = \frac{1}{\bar{x}} \exp(-x/\bar{x}), \tag{4.35}$$

respectively, where $\bar{x} = P_B L_{pbs} \Omega_x$, $\bar{z} = L_{sd} \Omega_z / N_0$, $\Omega_x = \Omega_h$, $\Omega_z = \Omega_g$. Substituting (4.34) and (4.35) in (4.28), taking $ax - bc = t$ and simplifying we have

$$F_\gamma(\gamma) = 1 - \frac{1}{a\bar{x}} \exp\left(-\frac{\gamma}{\bar{z}a}\right) \exp\left(-\frac{bc}{\bar{x}a}\right) \times \int_0^\infty \exp\left(-\frac{t}{\bar{x}a}\right) \exp\left(-\gamma \frac{bc+da}{t\bar{z}a}\right) dt. \quad (4.36)$$

Using [32, 3.471-9], the CDF is calculated as

$$F_\gamma(\gamma) = 1 - 2 \exp\left(-\frac{\gamma}{\bar{z}a}\right) \exp\left(-\frac{bc}{\bar{x}a}\right) \sqrt{\gamma F} K_1\left(2\sqrt{\gamma F}\right) \quad (4.37)$$

where $F = (bc + da)/a^2 \bar{x} \bar{z}$. Then, substituting (4.37) in (4.26) the throughput for the special case of Rayleigh fading is obtained.

4.5.2 Outage probability and throughput of NL-EH-2 model

Considering (4.4) and (4.11), the conditional CDF of the received SNR is given as

$$F_\gamma(\gamma) = F_Z\left(\gamma \frac{X+b}{aX} | X\right). \quad (4.38)$$

Averaging over (4.6), we have

$$F_\gamma(\gamma) = \int_0^\infty F_Z\left(\gamma \frac{X+b}{aX} | X\right) f_X(x) dx. \quad (4.39)$$

Substituting (4.5), and (4.6) in (4.39), and simplifying, we have

$$F_\gamma(\gamma) = 1 - \Theta(i, \gamma) \int_0^\infty \left(\frac{x+b}{ax}\right)^i x^{m_x-1} \times \exp(-x\phi_x - \phi_z \gamma b/ax) dx \quad (4.40)$$

where $\Theta(i, \gamma) = D \exp(-\phi_z \gamma/a) \sum_{i=0}^{m_z-1} (\gamma \phi_z)^i / i!$. Using [32, 1.111], we have

$$F_\gamma(\gamma) = 1 - \Theta(i, \gamma) \left(\frac{1}{a}\right)^i \sum_{k=0}^i \binom{i}{k} b^{i-k} \times \int_0^\infty x^{m_x-i+k-1} \exp(-x\phi_x - \phi_z \frac{\gamma b}{ax}) dx. \quad (4.41)$$

Applying [32, 3.471-9] and simplifying, we have

$$F_\gamma(\gamma) = 1 - 2D\xi(i, k) \gamma^{0.5(m_x+i+k)} \exp(-\phi_z \frac{\gamma}{a}) \times K_{m_x+k-i}\left(2\sqrt{\frac{\phi_x \phi_z b}{a} \gamma}\right) \quad (4.42)$$

where

$$\xi(i, k) = \sum_{i=0}^{m_z-1} \frac{(\phi_z/a)^i}{i!} \sum_{k=0}^i \binom{i}{k} b^{i-k} \left(\frac{\phi_z b}{\phi_x a} \right)^{(m_x+k-i)/2}. \quad (4.43)$$

Then, substituting (4.42) in (4.26) the system throughput for NL-EH-2 is obtained.

Special case of Rayleigh fading for NL-EH-2 model The outage probability and throughput of the considered system for NL-EH-2 model given in (4.11) is investigated under the assumption of Rayleigh fading for both links $PB \rightarrow S$ and $S \rightarrow D$. Applying (4.34) and (4.35) in (4.39), we have

$$F_\gamma(\gamma) = 1 - \frac{1}{\bar{x}} \exp(-\gamma/a\bar{z}) \int_0^\infty \exp\left(-\frac{x}{\bar{x}} - \frac{b\gamma}{ax\bar{z}}\right) dx. \quad (4.44)$$

Using [32, 3.471-9], the CDF of received SNR at node D is obtained as

$$F_\gamma(\gamma) = 1 - 2 \exp\left(-\frac{\gamma}{\bar{z}a}\right) \sqrt{\frac{b\gamma}{a\bar{z}\bar{x}}} K_1\left(2\sqrt{\frac{b\gamma}{a\bar{z}\bar{x}}}\right). \quad (4.45)$$

Finally, the throughput of considered system for NL-EH-2 is calculated by substituting (4.45) in (4.26).

4.5.3 Outage probability and throughput of linear and piece-wise linear models

Considering Table 4.1, the conditional CDF of the received SNR at D given in (4.4) for both linear and piece-wise linear EH models is written as

$$F_\gamma(\gamma|X) = \Pr(Z < \gamma/\eta X) = F_Z\left(\frac{\gamma}{\eta X}|X\right) \quad (4.46)$$

where $F_Z(z)$ is given in (4.5). For the linear EH model, the CDF is calculated as

$$F_\gamma^l(\gamma) = \int_0^\infty F_Z\left(\frac{\gamma}{\eta X}|X\right) f_X(x) dx. \quad (4.47)$$

Substituting (4.5) and (4.6) in (4.47) and applying [32, 3.471-9], the CDF in (4.47) is rewritten as

$$\begin{aligned} F_\gamma^l(\gamma) = & 1 - 2D \sum_{i=0}^{m_z-1} \frac{(\phi_z \gamma / \eta)^i}{i!} \left(\frac{\phi_z \gamma}{\eta \phi_x} \right)^{(m_x-i)/2} \\ & \times K_{m_x-i}\left(2\sqrt{\phi_z \gamma \phi_x / \eta}\right). \end{aligned} \quad (4.48)$$

For the piece-wise linear EH model given in Table 4.1, the CDF is expressed as

$$F_\gamma^{pw}(\gamma) = P(\eta X Z < \gamma, X < P_{th}) + P(\eta P_{th} Z < \gamma, X > P_{th}). \quad (4.49)$$

Considering (4.5), and (4.6), and simplifying we have

$$F_{\gamma}^{pw}(\gamma) = F_X(P_{th}) - E_2(\gamma) + F_Z\left(\frac{\gamma}{P_M}\right)(1 - F_X(P_{th})) \quad (4.50)$$

where $P_M = \eta P_{th}$. Here, $E_2(\gamma)$ is calculated by substituting (4.5) and (4.6) in (4.49) as

$$E_2(\gamma) = D \sum_{i=0}^{m_z-1} \frac{(\phi_z \gamma / \eta)^i}{i!} \int_0^{P_{th}} x^{m_x-i-1} e^{-x\phi_x} e^{-\phi_z \gamma / \eta x} dx. \quad (4.51)$$

Moreover, applying [32, 1.211-1], [40, 11], [32, 9.31-2] and, [40, 26], (4.51) is obtained as

$$E_2(\gamma) = D \sum_{i=0}^{m_z-1} \frac{(\phi_z \gamma / \eta)^i}{i!} \sum_{k=0}^N \frac{(-\phi_x)^k}{k!} P_{th}^{m_x-i+k} \times G_{21}^{02} \left(\frac{\eta P_{th}}{\gamma \phi_z} \middle| \begin{matrix} 1, 1 - (m_x - i + k) \\ -(m_x - i + k) \end{matrix} \right) \quad (4.52)$$

where $N \rightarrow \infty$, and $G_{pq}^{mn} \left(\cdot \middle| \begin{matrix} a_p \\ b_q \end{matrix} \right)$ is the Meijer-G function [32, 9.301]. Finally, the throughputs of linear and piece-wise linear EH models are obtained by substituting (4.48) and (4.50), respectively in (4.26).

Different from the NL-EH models previously given in the literature, we obtained closed-form outage probability expressions for the Nakagami- m channels in eqs. (4.32), and (4.42), and for the special case of Rayleigh fading channels in eqs. (4.37) and (4.45) for NL-EH-1 and NL-EH-2 models, respectively.

4.6 BEP Analyses

In this section, closed-form expressions for the BEP of the considered system for NL-EH-1, NL-EH-2, linear and piece-wise linear EH models are derived. The BEP is calculated as

$$P_b = \frac{A\sqrt{B}}{2\rho\sqrt{\pi}} \int_0^{\infty} \frac{e^{-B\gamma}}{\sqrt{\gamma}} F_{\gamma}(\gamma) d\gamma \quad (4.53)$$

where $F_{\gamma}(\gamma)$ is the CDF obtained in (4.32), (4.37), (4.42), and (4.45) for NL-EH-1 and NL-EH-2 in Nakagami- m and Rayleigh fading channels, respectively. Here, constants A and B are modulation specific parameters [33] and $\rho = \log_2 M$, M being the modulation order. Under the assumption of Gray mapping (4.53) provides a tight upper-bound at high and medium SNR regions. Substituting (4.32), (4.37), (4.42), and

(4.45) in (4.53), (4.53) is obtained in the general form of

$$P_b = \frac{A\sqrt{B}}{2\rho\sqrt{\pi}}(\Xi - I_\zeta) \quad (4.54)$$

where $\zeta \in \{1, 2, 3, 4\}$ and

$$\Xi = \int_0^\infty \frac{e^{-B\gamma}}{\sqrt{\gamma}} d\gamma = \sqrt{\frac{\pi}{B}}. \quad (4.55)$$

(4.55) is calculated using [32, 3.381-4]. In (4.54), I_1 , I_2 , I_3 , and I_4 are obtained by substituting (4.32), (4.37), (4.42), and (4.45), respectively, as explained in the following subsections.

4.6.1 BEP analysis for NL-EH-1 model

The BEP for the NL-EH-1 model given in (4.7) is calculated for Nakagami- m fading channel, from (4.54) by taking for $\zeta = 1$

$$I_1 = \varpi(i, k, l) \int_0^\infty \frac{e^{-B\gamma}}{\sqrt{\gamma}} \gamma^{0.5(u+i)} \exp\left(-\phi_z \frac{\gamma}{a^2}\right) \times K_{u-i}\left(2\sqrt{\frac{\phi_x \phi_z J}{a^2}} \gamma\right) d\gamma. \quad (4.56)$$

Furthermore, assuming $\gamma = z^2$ in (4.56), we have

$$I_1 = 2\varpi(i, k, l) \int_0^\infty z^{u+i} \exp(-z^2 \kappa) K_{u-i}(\delta z) dz \quad (4.57)$$

where $\kappa = B + \phi_z/a$ and $\delta = 2\sqrt{\phi_x \phi_z J/a^2}$. Using [32, 6.631-3] in (4.57) and simplifying we have

$$I_1 = \varpi(i, k, l) \left(\frac{1}{\sqrt{\kappa}}\right)^{\varsigma+i} \Gamma(0.5 + \varsigma) \Gamma(0.5 + i) \times \exp(\delta^2/8\kappa) W_{-0.5(\varsigma+i), 0.5(\varsigma-i)}(\delta^2/4\kappa)/\delta \quad (4.58)$$

where $\varsigma = k + l + 1$, $W_{\cdot, \cdot}(\cdot)$ is the Whittaker function [32, 9.22]. Finally, the closed-form BEP for NL-EH-1 model is calculated substituting (4.58) in (4.54).

Special case of Rayleigh fading for NL-EH-1 The BEP for Rayleigh fading channel is calculated replacing for $\zeta = 2$

$$I_2 = 2\sqrt{F} \exp\left(-\frac{bc}{\bar{x}a}\right) \int_0^\infty \exp\left(-\frac{\gamma}{\bar{x}a} - B\gamma\right) K_1\left(2\sqrt{\gamma F}\right) d\gamma. \quad (4.59)$$

in (4.54). Moreover, using [32, 6.614-4] in (4.59), we have

$$I_2 = \exp\left(-\frac{bc}{\bar{x}a} + \frac{F}{2\alpha}\right) \Gamma(1.5) \Gamma(0.5) W_{-0.5,0.5}\left(\frac{F}{\alpha}\right) / \sqrt{\alpha} \quad (4.60)$$

where $\alpha = B + 1/\bar{z}a$. Thus, the closed-form BEP expression for the NL-EH-1 model in Rayleigh fading is calculated by substituting (4.60) in (4.54).

4.6.2 BEP analysis for NL-EH-2 model

The BEP for the NL-EH-2 model given in (4.11) is calculated for Nakagami- m fading channel from (4.54) by replacing for $\zeta = 3$

$$I_3 = 2D\xi(i, k) \int_0^\infty \gamma^{0.5(m_x+k+i-1)} \exp\left(-\gamma\left(\frac{\phi_z}{a} + B\right)\right) \times K_{m_x+k-i}\left(2\sqrt{\frac{\phi_x\phi_z b}{a}}\gamma\right) d\gamma. \quad (4.61)$$

Assuming $\gamma = z^2$ in (4.61), we have

$$I_3 = 4D\xi(i, k) \int_0^\infty z^{m_x+k+i} \exp(-z^2\kappa) K_{m_x+k-i}(\sigma z) dz \quad (4.62)$$

where $\kappa = B + \phi_z/a$ and $\sigma = 2\sqrt{\phi_x\phi_z b/a}$. Using [32, 6.631-3] and simplifying, (4.62) is rewritten as

$$I_3 = 2D\xi(i, k) \left(\frac{1}{\sqrt{\kappa}}\right)^{m_x+k+i} \Gamma(0.5 + m_x + k) \Gamma(0.5 + i) \times \exp(\sigma^2/8\kappa) W_{-0.5(m_x+k+i), 0.5(m_x+k-i)}(\sigma^2/4\kappa) / \sigma. \quad (4.63)$$

Finally, substituting (4.63) in (4.54), the closed-form BEP expression for NL-EH-2 is obtained.

Special case of Rayleigh fading for NL-EH-2 The BEP is calculated from (4.54) by taking $\zeta = 4$ and replacing

$$I_4 = 2\sqrt{\frac{b}{a\bar{z}\bar{x}}} \int_0^\infty \exp\left(-\gamma\left(B + \frac{1}{a\bar{z}}\right)\right) K_1\left(2\sqrt{\frac{b}{a\bar{z}\bar{x}}}\gamma\right) d\gamma. \quad (4.64)$$

Using [32, 6.614-4], we have

$$I_4 = \exp\left(\frac{\beta}{2\alpha}\right) \Gamma(1.5) \Gamma(0.5) W_{-0.5,0.5}\left(\frac{\beta}{\alpha}\right) / \sqrt{\alpha} \quad (4.65)$$

where $\alpha = B + 1/\bar{z}a$ and $\beta = b/a\bar{z}\bar{x}$. Hence, the BEP expression for NL-EH-2 model under Rayleigh fading channel is derived substituting (4.65) in (4.54).

4.6.3 BEP analysis for linear and piece-wise linear EH models

The conditional BEP of the system considering both linear and piece-wise linear EH models can be expressed as

$$P_b(e|X) = \frac{A\sqrt{B}}{2\rho\sqrt{\pi}} \int_0^\infty \frac{e^{-B\gamma}}{\sqrt{\gamma}} F_\gamma(\gamma|X) d\gamma = \frac{A\sqrt{B}}{2\rho\sqrt{\pi}} (\Xi - M_2(x)) \quad (4.66)$$

where $F_\gamma(\gamma|x) = F_Z(\gamma/\eta X|X)$ is given in (4.46) and (4.5). Moreover, Ξ is obtained in (4.55), and

$$M_2(x) = \sum_{i=0}^{m_z-1} \frac{(\phi_z/\eta x)^i}{i!} \int_0^\infty \gamma^{i-0.5} \exp\left(-\gamma\left(B + \frac{\phi_z}{\eta x}\right)\right) d\gamma. \quad (4.67)$$

Applying [32, 3.381-4] in (4.67) and after some mathematical simplifications, we have

$$M_2(x) = \sqrt{\frac{\eta}{\phi_z}} \sum_{i=0}^{m_z-1} \frac{\Gamma(i+0.5)}{i!} \frac{\sqrt{x}}{(1+xG)^{i+0.5}} \quad (4.68)$$

where $G = \eta B/\phi_z$. For the linear EH model, the unconditional BEP of the considered system is calculated as

$$P_b^l = \int_0^\infty P_b(e|x) f_X(x) dx = \frac{A\sqrt{B}}{2\rho\sqrt{\pi}} (M_{11} - M_{21}) \quad (4.69)$$

where

$$M_{11} = \Xi \int_0^\infty f_X(x) dx = \Xi \quad (4.70)$$

and

$$M_{21} = \int_0^\infty M_2(x) f_X(x) dx \quad (4.71)$$

where $M_2(x)$ and $f_X(x)$ are given in (4.67) and (4.6), respectively. Substituting, simplifying, and using [40, Eq.10], we have,

$$\begin{aligned} M_{21} = & D \sqrt{\frac{\eta}{\phi_z}} \sum_{i=0}^{m_z-1} \frac{1}{i!} \int_0^\infty x^{m_x-0.5} \exp(-\phi_x x) \\ & \times G_{11}^{11} \left(xG \middle| \begin{matrix} 0.5-i \\ 0 \end{matrix} \right) dx. \end{aligned} \quad (4.72)$$

Using [32, 7.813-1], we have

$$\begin{aligned} M_{21} = & D \sqrt{\frac{\eta}{\phi_z}} \sum_{i=0}^{m_z-1} \frac{1}{i!} \phi_x^{-(m_x+0.5)} \\ & \times G_{21}^{12} \left(G/\phi_x \middle| \begin{matrix} -m_x+0.5, 0.5-i \\ 0 \end{matrix} \right) dx. \end{aligned} \quad (4.73)$$

Finally, the BEP expression for the linear model is obtained substituting (4.70) and (4.73) in (4.69).

Now, considering the piece-wise linear EH model, the BEP of the considered system is written as

$$P_b^{pw} = \underbrace{\int_0^{P_{th}} P_b(e|x) f_X(x) dx}_{T_1} + \underbrace{P_b(e|X = P_{th})}_{T_2} \underbrace{\int_{P_{th}}^{\infty} f_X(x) dx}_{T_3} \quad (4.74)$$

where T_2 is calculated by substituting P_{th} in (4.66), and $T_3 = 1 - F_X(P_{th})$ is calculated using (4.9). Moreover, we have

$$T_1 = \frac{A\sqrt{B}}{2\rho\sqrt{\pi}}(N_{11} - N_{21}) \quad (4.75)$$

where

$$N_{11} = \Xi \int_0^{P_{th}} f_X(x) dx = \Xi F_X(P_{th}) \quad (4.76)$$

and

$$N_{21} = \int_0^{P_{th}} M_2(x) f_X(x) dx \quad (4.77)$$

where $M_2(x)$ and $f_X(x)$ are given in (4.67) and (4.6), respectively. Substituting in (4.77), simplifying and applying $y = xG$, the upper bound of integration is changed from P_{th} to GP_{th} and (4.77) is rewritten as

$$N_{21} = \begin{cases} \Delta(i)(R_1 + R_2), & GP_{th} \geq 1 \\ \Delta(i)R_3, & GP_{th} < 1 \end{cases} \quad (4.78)$$

where

$$\Delta(i) = D \sqrt{\frac{\eta}{\phi_z}} \sum_{i=0}^{m_z-1} \frac{\Gamma(i+0.5)}{i! G^{m_x+0.5}}$$

and

$$R_l = \int \frac{y^{m_x-0.5}}{(1+y)^{i+0.5}} \exp(-\phi_x y/G) dy \quad (4.79)$$

where $\{0, 1\}$, $\{1, GP_{th}\}$, and $\{0, GP_{th}\}$ are the integration boundaries for $\iota = 1$, $\iota = 2$, and $\iota = 3$, respectively. Considering $\iota = 1$, and $\iota = 3$ in (4.79) and applying [40, 3], we have

$$R_l = \sum_{k=0}^N (-\epsilon)_k \frac{(-1)^k}{k!} \int_0^v y^{m_x-0.5+k} \exp(-\phi_x y/G) dy \quad (4.80)$$

where $\varepsilon = -(i + 0.5)$, and $v = 1$ and GP_{th} for R_1 and R_3 , respectively. Applying [38, 07.20.07.0001.01] in (4.80) we have

$$R_1 = \sum_{k=0}^N (-\varepsilon)_k \frac{(-1)^k \Gamma(m_x + k + 0.5)}{k! \Gamma(m_x + k + 1.5)} \times {}_1F_1\left(m_x + k + 0.5; m_x + k + 1.5; -\frac{\phi_x}{G}\right), \quad (4.81)$$

and

$$R_3 = \sum_{k=0}^N (-\varepsilon)_k \frac{(-1)^k \Gamma(m_x + k + 0.5)}{k! \Gamma(m_x + k + 1.5)} (P_{th}G)^{m_x + k + 0.5} \times {}_1F_1(m_x + k + 0.5; m_x + k + 1.5; -\phi_x P_{th}) \quad (4.82)$$

where ${}_1F_1(a; b; \cdot)$ is the Kummer confluent hypergeometric function given in [38, 07.20.02.0001.01]. Considering $\iota = 2$ in (4.79) and applying [40, Eq.3], we have

$$R_2 = \sum_{k=0}^N (-\varepsilon)_k \frac{(-1)^k}{k!} (R_{21} - R_{22}) \quad (4.83)$$

where

$$R_{21} = \int_1^\infty y^{m_x - i - k - 1} \exp(-\phi_x y / G) dy \\ = \left(\frac{\phi_x}{G}\right)^{-(m_x - i - k)} \Gamma(m_x - i - k, \phi_x / G), \quad (4.84)$$

and

$$R_{22} = \int_{P_{th}G}^\infty y^{m_x - i - k - 1} \exp(-\phi_x y / G) dy \\ = \left(\frac{\phi_x}{G}\right)^{-(m_x - i - k)} \Gamma(m_x - i - k, \phi_x P_{th}). \quad (4.85)$$

Here, (4.84) and (4.85) are calculated using [32, 3.351-2]. Then, (4.83) is calculated by substituting (4.84) and (4.85). Moreover, (4.78) is calculated using (4.81), (4.82), and (4.83). Finally, substituting (4.75), T_2 and T_3 in (4.66), and (4.9), respectively, the BEP expression for the piece-wise linear EH model is obtained.

Thanks to the mathematical tractability of our NL-EH models, we obtained closed-form expressions for the average BEP of the considered power-constraint system. Moreover, to gain a better insight through the new NL-EH models, we calculated BEP of the reference systems applying linear and piece-wise linear EH models. The detailed effects of the system parameters on the obtained BEP expressions are evaluated in the next section.

Table 4.5 : Fitting parameters of different EH models given in Table 4.1 with data measurements taken from energy harvesting circuits of [1], and [2].

	[1]	[2]
EH models	Parameters	Parameters
Linear [14]	$\eta = 0.6$	$\eta = 0.25$
Piece-wise linear [8]	$\eta = 0.6,$ $P_{th} = 410$ $a = 0.00409,$	$\eta = 0.25,$ $P_{th} = 29.5$ $a = 0.08271,$
NL-EH [6]	$b = -63.88202,$ $M = 244$ $p_0 = 406.5963,$ $p_1 = -46878.9296,$ $p_2 = 3389602.655,$	$b = 8.34322,$ $M = 7.3$ $p_0 = 10.24347,$ $p_1 = -25.14551,$ $p_2 = 16.37787$
NL-EH [7]	$q_1 = 570.147,$ $q_2 = -115371.7,$ $q_3 = 1.0291e7$	$q_1 = 21.21917,$ $q_2 = -7.88436,$ $q_3 = 138.76844$

4.7 Performance Evaluation

In this section, simulation and theoretical results are obtained for different values of the system parameters. In all figures of this section, lines denote the theoretical results while symbols or symbols with dashed lines present the simulation results. In addition to two the proposed NL-EH models, the simulation results for linear, piece-wise linear, and existing NL-EH models are provided which gives an extra insight in analyzing the system performance. In all cases, the theoretical results are in perfect match with the Monte-Carlo simulation results. Theoretical curves for average harvested power are obtained from (4.14), (4.20), (4.21), and (4.22). Theoretical throughput curves are depicted by replacing (4.32), (4.42), (4.48), and (4.50) in (4.26). For the BER analysis, theoretical results are obtained by applying (4.54), (4.69), and (4.74). The fitting parameters for [2], [1] are listed in Table 4.3 and Table 4.4, for the NL-EH-1 and NL-EH-2 models, respectively, while the coefficient of other models are provided in Table 4.5. Unless stated differently, the fading parameters for each link are taken as $m_h = m_g = 2$ and, the nodes S and D are both equipped with $N_s = N_d = 2$ receive

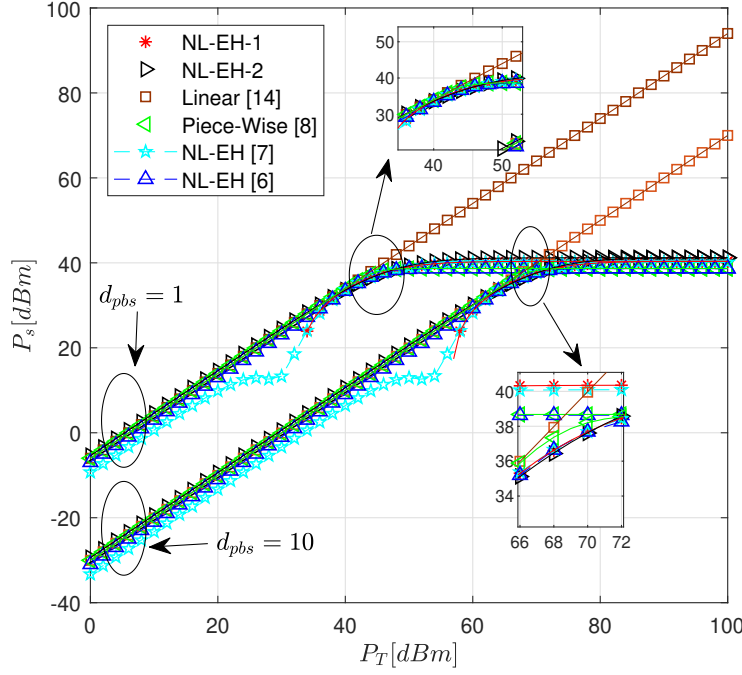


Figure 4.6 : Average harvested power versus P_T where data measurements are taken from [2].

antennas. It is assumed that the modulation is binary phase-shift keying (BPSK) and the system throughput is obtained for $R = 1$ bits/sec/Hz. Furthermore, it is assumed that $d_{pbs} = d_{sd} = 1$ and the path-loss coefficient is taken equal to $\nu = 2.7$ [41]. Moreover, the channel power gains for both links are chosen as $\Omega_h = \Omega_g = 1$, and the noise power at node D is fixed to $N_0 = 1$. Note that the optimization of the time slots dedicated for EH, and IP is a difficult task due to the depending parameters P_T , d_{pbs} , d_{sd} , ν , N_s , N_d , m_h , m_g , Ω_h , Ω_g , R .

Figure 4.6 describes the average harvested power at node S versus P_T for data measurements taken from (4.14). The resulting curves are depicted for $d_{pbs} = 1$, and $d_{pbs} = 10$. It is shown that the amount of average harvested power for $d_{pbs} = 1$ is 0 dBm which is approximately 25 dBm more compared to the case of $d_{pbs} = 10$. This verifies that the distance d_{pbs} has a significant effect on the amount of harvested power. Moreover, it is seen that, the average harvested powers of the NL-EH models for $d_{pbs} = 1$, and $d_{pbs} = 10$ are saturated at $P_T = 45$ dBm and $P_T = 70$ dBm, respectively. Please note that all NL-EH models are saturated at a pre-defined P_{th} value. Furthermore, compared to $d_{pbs} = 1$, 25 dBm more power is needed to saturate for $d_{pbs} = 10$.

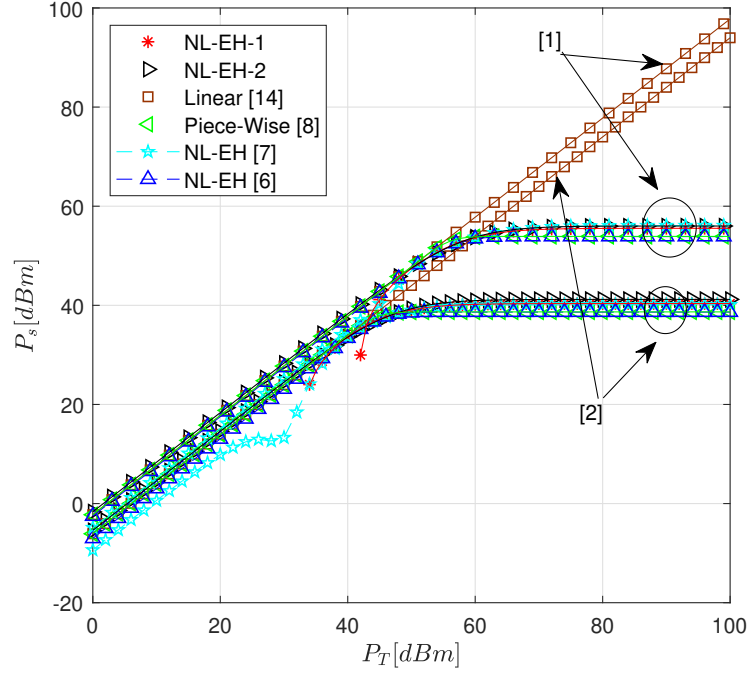


Figure 4.7 : Average harvested power versus P_T where $d_{pbs} = 1$, data measurements are taken from [1] and [2].

The behavior of average harvested power versus P_T is shown in Figure 4.7 considering the same system parameters for two different EH circuits given in [2] and [1] for $d_{pbs} = 1$. It is observed that the EH circuit in [1] provides approximately 2 dBm more average harvested power compared to that in [2] considering the linear EH model. Moreover, it is shown that the average harvested power saturates at $P_T = 40$ dBm, and $P_T = 55$ dBm for the EH circuits given in [2] and [1], respectively and results from the design of EH system. This means that the EH circuit design has an important impact on the performance of EH system since more power is harvested at power-constraint node S. However, despite the EH circuit design, NL-EH models saturate at a predefined threshold value. Please note that all the NL-EH models along with the proposed NL-EH models have the same behavior on the amount of average harvested power.

In Figure 4.8 average harvested power is depicted versus N_s . The results are provided for $d_{pbs} = 1$ and $P_T = 50$ dBm. It is concluded from Figure 4.8 that more antennas at node S result in higher average harvested power at the EH node which significantly affects the proposed system performance. However, it is shown that by increasing the number of receive antennas at node S, average harvested power approximately

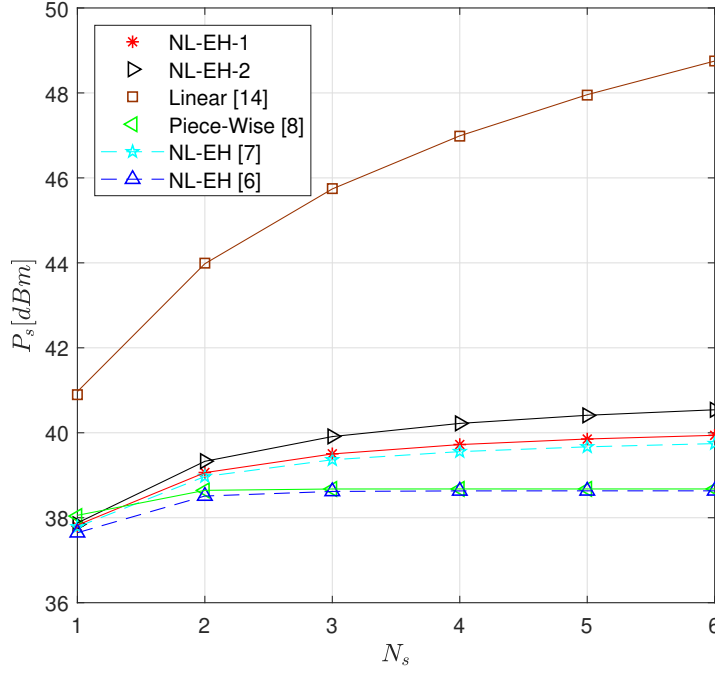


Figure 4.8 : Average harvested power versus N_s for $P_T = 50$ dBm and $d_{pbs} = 1$.

saturates to the values between 38 dBm and 41 dBm for different NL-EH models when $N_s = 6$. In other words, node S with $N_s = 3$ antennas provides almost equal amount of harvested power compared to node S with $N_s = 6$ antennas. Please note that the differences among NL-EH models are related to the curve fitting of the experimental measured data in a specified interval of input values. Moreover, it is seen from Figure 4.8 that the amount of average harvested power for the linear EH model is overestimated compared to the NL-EH and piece-wise linear EH models where for $N_s = 6$ the average harvested power for the linear EH model is 48.5 dBm. This means that linear EH model overestimates approximately 8.5 dBm more average harvested power compared to NL-EH models.

In Figure 4.9, the system throughput is depicted versus R . The results are provided for different numbers of receive antennas N_s , and N_d . It is seen that by increasing the number of antennas at both nodes S and D, the throughput is increased. Besides, it is observed from Figure 4.9 that all of the throughput curves start from approximately $\tau = 0.2$, reach their maximum values and then get decreased by increasing the value of R until they get toward zero. Moreover, it is observed that the linear EH model overestimates the system performance for all cases. Additionally, the more receive

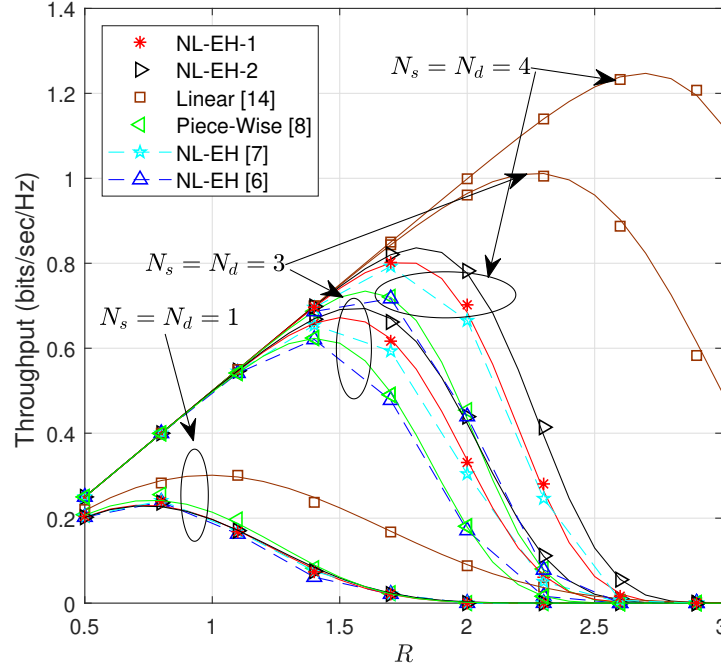


Figure 4.9 : Throughput of considered system versus R (bits/sec/Hz) for $P_T = 50$ dBm, $d_{pbs} = d_{sd} = 1$.

antennas at nodes S and D, the more overestimation of throughput is resulted. In other words, the differences between NL-EH models and linear EH model throughput curves are increased by increasing N_s and N_d . This is due to the fact that no saturation power is defined in linear EH model.

Figure 4.10 presents BER versus P_T when $d_{pbs} = d_{sd} = 1$. The curves are obtained for $N_s = N_d = 2$. It is shown that at high SNR values the system performance considering the NL-EH and piece-wise linear EH models go to the error floor while the linear EH model system performance is increased at high SNR values due to the overestimation of harvested power which causes an unreasonable expectation of the system performance. However, at low SNR values, the BER performances of linear/piece-wise linear/NL-EH models are equivalent. Additionally, based on the obtained parameters from curve fitting of [2], piece-wise linear and Logistic EH models provide approximately the same performance. Similarly, the heuristic model, fraction model, and the proposed NL-EH-1 and NL-EH-2 models provide similar system performances. This comes from the fact that the difference in these results is caused by curve fitting functions which are calculated based on a specific interval of data measurements.

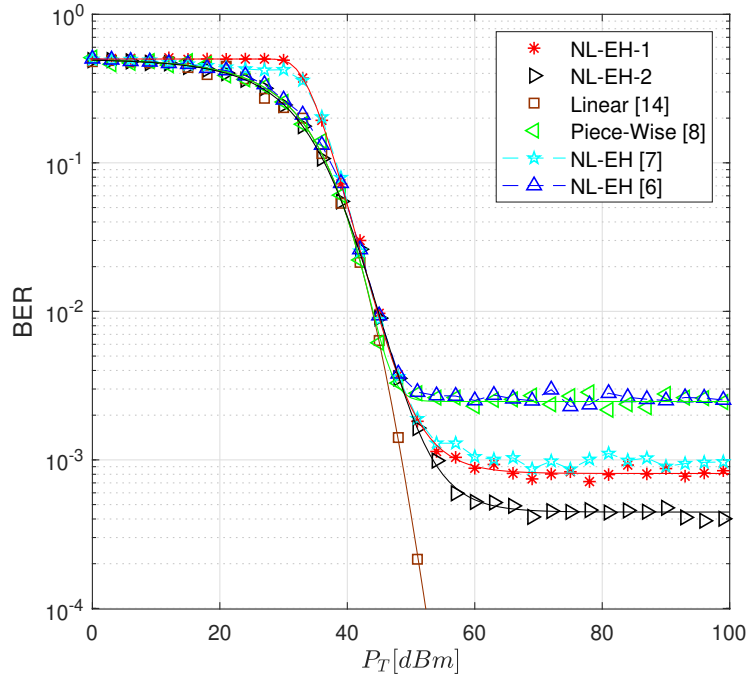


Figure 4.10 : BER versus P_T for $N_s = N_d = 2$, $d_{pbs} = d_{sd} = 1$.

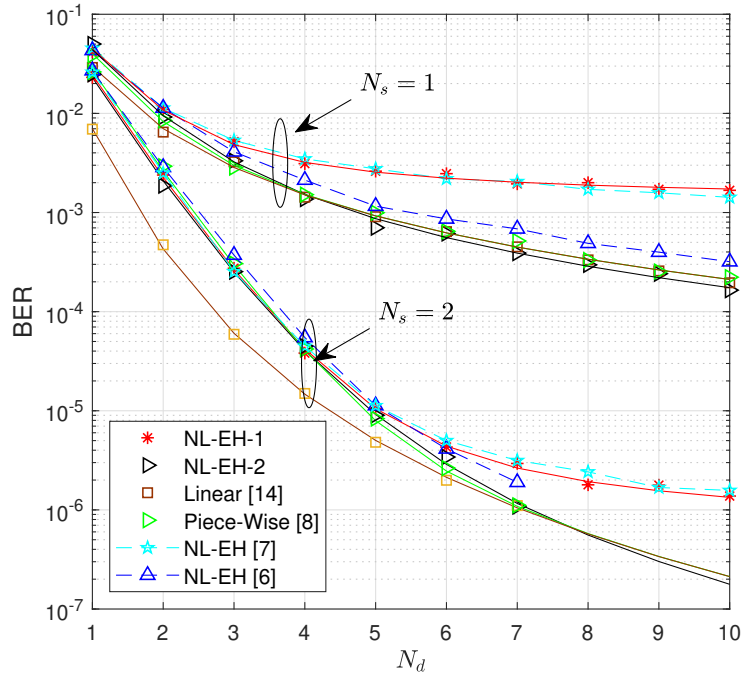


Figure 4.11 : BER versus N_d for $P_T = 50$ dBm, $d_{pbs} = d_{sd} = 1$ where the results are provided for $N_s = 1$ and $N_s = 2$.

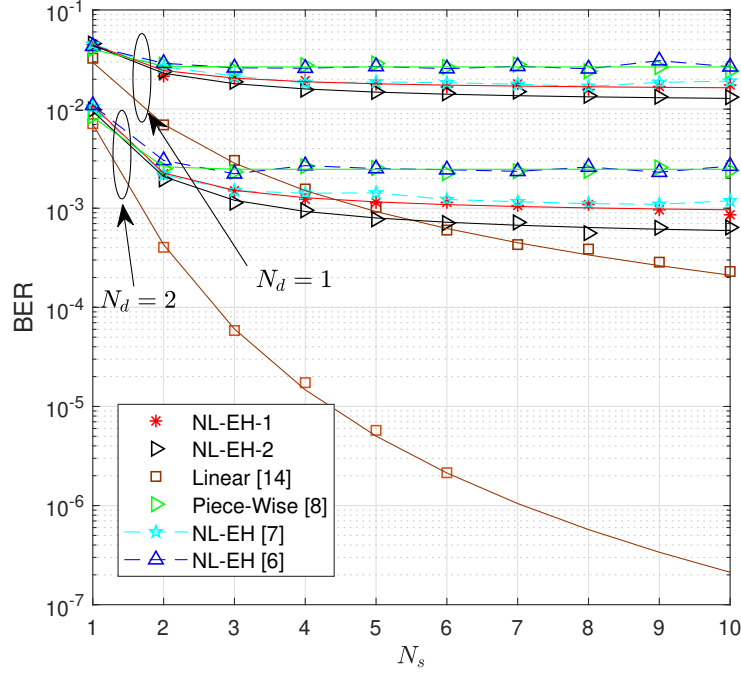


Figure 4.12 : BER versus N_s for $P_T = 50$ dBm, $d_{pbs} = d_{sd} = 1$ where the results are provided for $N_d = 1$ to $N_d = 2$.

The BER performance curves versus N_d are shown in Figure 4.11. Here, the results are provided for $P_T = 50$ dBm, $d_{pbs} = d_{sd} = 1$, $N_s = 1$ and $N_s = 2$. It is seen that by increasing N_d the BER performance gets better for both N_s values. Also, the linear EH model outperforms the piece-wise linear and NL-EH models for $N_s = 1$ and $N_s = 2$ due to more harvested power. It comes from the fact that more receive antennas at node S provides more harvested energy which results in higher transmit power at node S and consequently high received SNR at node D. Hence, it ensures a better system performance.

In Figure 4.12, the BER performance versus N_s is given for $P_T = 50$ dBm, and $d_{pbs} = d_{sd} = 1$. It is observed that by increasing the number of antennas from $N_d = 1$ to $N_d = 2$ the system performance gets significantly better. Additionally, by increasing the number of antennas at S, more power is harvested which results in better system performance for the linear EH model. However, for the NL-EH and piece-wise linear EH models, at high input power, the harvested power is saturated at a threshold power value, and then even by increasing the number of antennas at S, namely N_s , the system performance goes to the error floor. Obviously, for both $N_d = 1$, and $N_d = 2$, the same performance are obtained compared to $N_s = 2$, and $N_s = 10$. In other words, the S node

with only $N_s = 1$ antenna provides approximately equivalent performance compared to $N_s = 10$. Considering the results from both Figure 4.11 and 4.12, it is concluded that the performances of piece-wise linear and NL-EH models are mostly dependent on the value of N_d rather than N_s since N_s provides only an increase in the harvested power which is saturated at some value. However, even with small harvested power values but more antennas at node D, the system performance is improved. However, for the linear EH model, since the system performance is directly related to N_s , increasing N_s , results in better system performance.

4.8 Conclusion

The performance of a wireless-powered communication system equipped with multiple antennas at both nodes S and D has been investigated based on the practical NL-EH models. Two new mathematically tractable NL-EH models have been proposed for which the PDFs and CDFs of the harvested power and their histograms have been provided considering the data measurements taken from the literature. Different from the NL-EH models given in the literature which are not analytically tractable, these new NL-EH models make the performance analyses analytically tractable. Furthermore, considering the Nakagami- m fading and as a special case the Rayleigh fading, closed-form expressions for the average harvested powers, outage probabilities, throughputs, and BER performances have been analytically derived. Moreover, for comparison purposes, the same performance measures have been analytically derived for linear and piece-wise linear EH models. In each case, theoretical derivations have been supported by Monte-Carlo simulation results and have been compared with the existent EH models given in the literature such as linear, piece-wise linear, and other NL-EH models. It has been shown that apart from linear EH model which overestimates the realistic system performance, the NL-EH models and piece-wise linear EH model provide practical system performances. Finally, the results obtained from different NL-EH models provide a comprehensive insight into the performance of the considered system. We have considered the case of one transmit antenna in this chapter, however it is possible to have more antennas for transmission at the source and even analyze the system performance based on transmit antenna

selection (TAS). However, these assumptions are out of scope of our chapter and left as a future work.





5. CONCLUSIONS

RF EH provides an environmental-friendly approach to the energy scarcity of next-generation wireless communication where the power-constraint nodes are empowered by scavenging ambient or dedicated power. Hence, this technology can extend the life time of information transmission among the IoT and sensors, which require a low amount of energy.

In this thesis, we have covered three different system models, and the performance of each considered system has been obtained for different EH models. All the parameters that affect the amount of harvested energy have been discussed, including the EH model type, distance, etc. For all the considered systems, the optimum system performance based on the system parameters has also been determined. Moreover, in the last part of this dissertation, we have proposed two mathematically tractable NL EH models for which the system performances have been obtained and compared for different system parameters and available EH models. Therefore, this has provided a comprehensive insight into the EH system design.

In the first part of the thesis, we have studied PU and SU BER performance in a CR network using an overlay spectrum sharing protocol. Here, SU cooperates with PU and in return it takes advantage of the PU licensed spectrum. Moreover, ST is assumed to be a power-constrained node that harvests RF power from PT and SR when SR operates in FD mode. The system analysis is also more realistic since an interference link between ST and PR has been taken into consideration. Closed-form expressions for the BEP of the overlay CR network under consideration have been obtained for both PU and SU. Moreover, Monte-Carlo computer simulation results for various system parameters and linear and NL EH models have been validated the theoretical derivations. For comparison purposes, the performance of PU without SU (direct transmission) has also been obtained, which brings up the improvement provided by the considered system. Furthermore, the system performances have been obtained with regard to a power allocation coefficient at SR to obtain the optimum

compromise between EH and information relaying. It has been shown that the linear EH model provides an accurate system performance only in low input power at the power-constraint node, while at high input power, the amount of harvested power is beyond the realistic values. Therefore, the obtained results mislead the researcher about the real performance of the EH system.

In the second part of the thesis, a comprehensive performance analysis of a WP DH AF relaying system has been performed, considering both linear and NL EH models. Source and relay have been considered to be battery-less and supplied by a dedicated PB in the investigated system. Results have been obtained for a comprehensive study taking into account both the piece-wise linear and linear EH models, which are typically assumed in the literature. Analytical results for the BEP and throughput of the system for both linear and piece-wise linear EH models have been compared with the results of Monte-Carlo simulations, which validate our theoretical derivations. It has been revealed that the linear EH model only yields realistic results at low levels of harvested energy and misrepresents the system performance at high levels of harvested energy. Consequently, this leads to a misunderstanding of the EH system design process. A comprehensive evaluation of the EH system architecture has been provided by the studied linear piece-wise EH model, which offers a realistic result for the system performance at both low and high harvested energy levels. In this part, we have considered that all nodes are equipped with a single antenna. However, it is feasible to add more antennas to the source, relay, and destination nodes and evaluate how the system performance improves by choosing the TAS and/or MRC techniques. In the third part of the thesis, based on the practical NL-EH models, the performance of a WPC system with several antennas at both nodes S and D has been studied. Two novel mathematically tractable NL-EH models have been proposed, for which the PDFs and CDFs of the harvested power, as well as their histograms, have been provided using data measurements from the literature. In contrast to the NL-EH models described in the literature, which are not analytically tractable, these novel NL-EH models enable analytical tractability of performance studies. Moreover, considering the Nakagami- m fading and, as a special case, the Rayleigh fading, closed-form expressions have been analytically derived for the average harvested powers, outage probabilities, throughput, and BER performances. The

same performance metrics have also been analytically derived for piece-wise linear and linear EH models for comparison purposes. Theoretical derivations in each case have been validated by Monte-Carlo simulation results and compared with the existing EH models, such as the linear, piece-wise linear, and other NL-EH models. It has been demonstrated that, in contrast to the linear EH model, which overestimates the performance of a realistic system, NL-EH models and piece-wise linear EH models provide accurate system performance. Finally, the results of various NL-EH models provide a deep understanding of the power-constraint systems' operation.

Consequently, in this thesis, we have analyzed three different system models with various EH models, which provide insight into the EH system designs. Moreover, two practical NL EH models have been proposed, which are well approximated the real experimental results. Our future work is devoted to the development and investigation of other EH system considering the proposed practical NL-EH models, which help researchers to gain comprehensive insight through this promising technology.



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APPENDICES

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APPENDIX A: PDF of $Z = X + Y$

The PDF of Z from (2.4) where for simplicity we assume $Z = P_h$ can be represented as [37, 6-43]

$$f_Z(z) = \int_0^z f_X(z-y)f_Y(y)dy \quad (\text{A.1})$$

where $X = \eta P_{pt} L_{41} |h_{41}|^2$ and $Y = \eta P_{sr_1} L_{43} |h_{43}|^2$ are exponentially and gamma distributed, respectively. The PDFs of X and Y are described as $f_X(x) = \psi_{41} \exp(-x\psi_{41})$ and $f_Y(y) = \psi_{43}^{m_{43}} y^{m_{43}-1} \exp(-\psi_{43}y) / \Gamma(m_{43})$ with $\psi_{41} = 1/\bar{x}$, $\psi_{43} = m_{43}/\bar{y}$, $\bar{x} = \eta P_{pt} L_{41} \Omega_{41}$ and $\bar{y} = \eta P_{sr_1} L_{43} \Omega_{43}$. By using [32, 3.381-1] and after some simplification, the PDF of Z can be written as

$$f_Z(z) = \alpha \exp(-\psi_{41}z) \beta^{-m_{43}} \gamma(m_{43}, \beta z) \quad (\text{A.2})$$

where $\alpha = \psi_{41} \psi_{43}^{m_{43}} / \Gamma(m_{43})$ and $\beta = \psi_{43} - \psi_{41}$. Then, applying [32, 8.352-6] to (A.2), the PDF of harvested power at ST (P_{st}) can be determined as

$$f_Z(z) = \rho \exp(-\psi_{41}z) - \rho \exp(-\psi_{43}z) \sum_{j=0}^{m_{43}-1} \frac{\beta^j}{j!} z^j \quad (\text{A.3})$$

where $\rho = \alpha \Gamma(m_{43}) / \beta^{m_{43}}$.



APPENDIX B: CDF of $\gamma_{23} = R/(1+Q)$

The CDF of γ_{23} in (2.6) is calculated as [37, 6-60]

$$\begin{aligned} F_{\gamma_{23}}(\gamma) &= \int_0^\infty F_R(\gamma(1+q))f_Q(q)dq \\ &= \int_0^\infty [1 - \exp(-\frac{\gamma(1+q)}{\bar{R}})]f_Q(q)dq \end{aligned} \quad (\text{B.1})$$

where $Q = \bar{w}Z$, $Z = X + Y$ and $f_Q(q) = f_Z(q/\bar{w})/\bar{w}$ where $f_Z(z)$ is calculated in (A.3) and $\bar{w} = L_{24}\Omega_{24}/\sigma^2$. $F_R(r)$ is the CDF of received SNR at PR from the link SR→PR which is exponentially distributed with $\bar{R} = P_{sr2}L_{23}\Omega_{23}/\sigma^2$. Substituting and simplifying, (B.1) is expressed as

$$F_{\gamma_{23}}(\gamma) = 1 - \exp(-\frac{\gamma}{\bar{R}})[A_1C - A_2(j)D] \quad (\text{B.2})$$

where $A_1 = \rho/\bar{w}$, $A_2(j) = A_1 \sum_{j=0}^{m_{43}-1} \frac{(\beta/\bar{w})^j}{j!}$. In (B.2),

$$C = \int_0^\infty \exp(-q(\frac{\gamma}{\bar{R}} + \nabla))dq = 1/(\frac{\gamma}{\bar{R}} + \nabla), \quad (\text{B.3})$$

$$D = \int_0^\infty q^j \exp(-q(\frac{\gamma}{\bar{R}} + \xi))dq = \Gamma(j+1)/(\frac{\gamma}{\bar{R}} + \xi)^{(j+1)} \quad (\text{B.4})$$

where $\nabla = \psi_{41}/\bar{w}$ and $\xi = \psi_{43}/\bar{w}$ for which [32, 3.310] and [32, 3.326-2] are applied. The CDF of received SNR at PR from the link SR→PR under an interference link from ST is calculated by substituting (B.3) and (B.4) in (B.2).



APPENDIX C: PDF of $\gamma_{23} = R/(1+Q)$

The PDF $f_{\gamma_{23}}(\gamma)$ in (2.6) is calculated as [37, 6-60]

$$f_{\gamma_{23}}(\gamma) = \int_0^\infty f_R(\gamma(1+q))f_Q(q)dq + \int_0^\infty qf_R(\gamma(1+q))f_Q(q)dq \quad (C.1)$$

where $f_Q(q) = A_1 \exp(-\nabla q) - A_2(j) \exp(-\xi q)q^j$, $f_R(r) = \exp(-r/\bar{R})/\bar{R}$. After substituting and some mathematical simplification, using [32, 3.326-2] we have

$$f_{\gamma_{23}}(\gamma) = (L_1 + L_2) \exp(-\gamma/\bar{R})/\bar{R} \quad (C.2)$$

where

$$L_1 = \left[\frac{A_1}{\nabla + \gamma/\bar{R}} - \frac{A_2(j)\Gamma(j+1)}{(\xi + \gamma/\bar{R})^{j+1}} \right], \quad (C.3)$$

$$L_2 = \left[\frac{A_1\Gamma(2)}{(\nabla + \gamma/\bar{R})^2} - \frac{A_2(j)\Gamma(j+2)}{(\xi + \gamma/\bar{R})^{j+2}} \right], \quad (C.4)$$

By re-expressing (C.2) as a Meijer-G function [39], we have $L_1 = L_{11} - L_{12}$ and $L_2 = L_{21} - L_{22}$ where

$$\begin{aligned} L_{11} &= \frac{A_1}{\nabla} G_{11}^{11} \left(\frac{\gamma}{\nabla \bar{R}} \middle| \begin{matrix} 0 \\ 0 \end{matrix} \right), \\ L_{12} &= \frac{A_2(j)}{\xi^{j+1}} G_{11}^{11} \left(\frac{\gamma}{\xi \bar{R}} \middle| \begin{matrix} -j \\ 0 \end{matrix} \right), \\ L_{21} &= \frac{A_1}{\nabla^2} G_{11}^{11} \left(\frac{\gamma}{\nabla \bar{R}} \middle| \begin{matrix} -1 \\ 0 \end{matrix} \right), \\ L_{22} &= \frac{A_2(j)}{\xi^{j+2}} G_{11}^{11} \left(\frac{\gamma}{\xi \bar{R}} \middle| \begin{matrix} -j-1 \\ 0 \end{matrix} \right). \end{aligned}$$

Substituting (C.3) and (C.4) in (C.2), $f_{\gamma_{23}}(\gamma)$ can be simply obtained.



APPENDIX D: Calculation of $T(\varepsilon)$

The integral $T(\varepsilon)$ in (2.24) can be written as

$$T(\varepsilon) = \int_0^\varepsilon \exp(\gamma/\bar{\gamma}_1) f_{\gamma_{23}}(\gamma) d\gamma = G_1(\varepsilon) + G_2(\varepsilon) \quad (\text{D.1})$$

where $f_{\gamma_{23}}(\gamma)$ is calculated in (C.2) and

$$G_1(\varepsilon) = \frac{1}{\bar{R}} \int_0^\varepsilon \exp(\gamma\phi) (L_{11} - L_{12}) d\gamma, \quad (\text{D.2})$$

$$G_2(\varepsilon) = \frac{1}{\bar{R}} \int_0^\varepsilon \exp(\gamma\phi) (L_{21} - L_{22}) d\gamma \quad (\text{D.3})$$

where $\phi = 1/\bar{\gamma}_1 - 1/\bar{R}$. Using [32, 1.211-1]

$$G_1(\varepsilon) = \frac{1}{\bar{R}} \sum_{k=0}^{\infty} \frac{\phi^k}{k!} \int_0^\varepsilon \gamma^k (L_{11} - L_{12}) d\gamma, \quad (\text{D.4})$$

$$G_2(\varepsilon) = \frac{1}{\bar{R}} \sum_{k=0}^{\infty} \frac{\phi^k}{k!} \int_0^\varepsilon \gamma^k (L_{21} - L_{22}) d\gamma. \quad (\text{D.5})$$

Substituting, applying some mathematical simplification and using [40, 26] we have

$$G_1(\varepsilon) = \frac{1}{\bar{R}} \sum_{k=0}^{\infty} \frac{\phi^k}{k!} (G_{11}(\varepsilon) - G_{12}(\varepsilon)), \quad (\text{D.6})$$

$$G_2(\varepsilon) = \frac{1}{\bar{R}} \sum_{k=0}^{\infty} \frac{\phi^k}{k!} (G_{21}(\varepsilon) - G_{22}(\varepsilon)) \quad (\text{D.7})$$

where

$$\begin{aligned} G_{11}(\varepsilon) &= \frac{A_1}{\nabla} \varepsilon^{k+1} G_{22}^{12} \left(\frac{\varepsilon}{\nabla \bar{R}} \middle| 0, -k \right), \\ G_{12}(\varepsilon) &= \frac{A_2(j)}{\xi^{j+1}} \varepsilon^{k+1} G_{22}^{12} \left(\frac{\varepsilon}{\xi \bar{R}} \middle| -j, -k \right), \\ G_{21}(\varepsilon) &= \frac{A_1}{\nabla^2} \varepsilon^{k+1} G_{22}^{12} \left(\frac{\varepsilon}{\nabla \bar{R}} \middle| -1, -k \right), \\ G_{22}(\varepsilon) &= \frac{A_2(j)}{\xi^{j+2}} \varepsilon^{k+1} G_{22}^{12} \left(\frac{\varepsilon}{\xi \bar{R}} \middle| -j-1, -k \right). \end{aligned}$$

Substituting (D.6) and (D.7) in (D.1), $T(\varepsilon)$ can be calculated.



APPENDIX E: Battery Threshold

The probabilities that Z given in Appendix A being below and over the threshold P_{th} is calculated as

$$\int_0^{\varsigma} f_Z(z) dz = \frac{\rho}{\psi_{41}} [1 - e^{-\varsigma \psi_{41}}] - \rho \sum_{j=0}^{m_{43}-1} \frac{\beta^j}{j!} \times [\psi_{43}^{-(j+1)} \gamma(j+1, \varsigma \psi_{43})] \quad (\text{E.1})$$

and

$$\int_{\varsigma}^{\infty} f_Z(z) dz = \frac{\rho}{\psi_{41}} [e^{-\varsigma \psi_{41}}] - \rho \sum_{j=0}^{m_{43}-1} \frac{\beta^j}{j!} \times [\psi_{43}^{-(j+1)} \Gamma(j+1, \varsigma \psi_{43})] \quad (\text{E.2})$$

respectively. (E.1) and (E.2) are obtained using [32, 3.381-1] and [32, 3.381-3], respectively.



APPENDIX F: Average Power

The average harvested power at the ST battery for the nonlinear EH model is calculated as

$$P_{ave}^{nL} = \int_0^{P_{th}} z f_Z(z) dz + P_{th} \int_{P_{th}}^{\infty} f_Z(z) dz - \int_0^{P_{min}} z f_Z(z) dz \quad (F.1)$$

where using [32, 3.381-1] and [32, 3.381-3], for the first and third integrals at the right-hand side of (F.1), we have

$$\begin{aligned} \int_0^{\varsigma} z f_Z(z) dz &= \frac{\rho}{\psi_{41}^2} \gamma(2, \varsigma \psi_{41}) - \rho \sum_{j=0}^{m_{43}-1} \frac{\beta^j}{j!} \\ &\times \left[\psi_{43}^{-(j+2)} \gamma(j+2, \varsigma \psi_{43}) \right] \end{aligned} \quad (F.2)$$

where $\varsigma \in \{P_{th}, P_{min}\}$. The second integral in (F.1) is obtained in (E.2). For the case of linear EH model, the average harvested power at the ST battery is calculated as [32, 3.326-2]

$$P_{ave}^L = \int_0^{\infty} z f_Z(z) dz = \frac{\rho}{\psi_{41}^2} - \rho \sum_{j=0}^{m_{43}-1} \frac{\beta^j}{j!} \frac{\Gamma(j+2)}{\psi_{43}^{j+2}}. \quad (F.3)$$



APPENDIX G: Calculation of $A(w)$

In (3.16), substituting (3.14), $A(w)$ is re-expressed as

$$A(w) = \frac{a}{2} \int_0^{P_{th}} \left[1 - \frac{1}{\sqrt{C + \frac{B}{z}}} \right] f_Z(z) dz, \quad (G.1)$$

where $C = (1 + w\bar{y})/w\bar{y}$, $B = 1/\bar{x}$ and $P_{th} = P_{th_s} = P_{th_r}$ and $f_Z(z)$ is given in (3.21). Substituting (3.21), (G.1) is written as

$$A(w) = \frac{a}{2} (E_1 - F_1(w)), \quad (G.2)$$

where

$$E_1 = \int_0^{P_{th}} f_Z(z) dz = D\gamma(m_z, P_{th}\phi_z)/\phi_z^{m_z} \quad (G.3)$$

and after simplifying,

$$F_1(w) = \frac{D}{\sqrt{B}} \int_0^{P_{th}} \frac{z^{m_z-0.5}}{\sqrt{1+z/G}} \exp(-z\phi_z) dz, \quad (G.4)$$

where $G = B/C$. Moreover, (G.3) is calculated using [32, 3.351-1]. Taking $z/G = x$ and after some mathematical modifications, (G.4) is given as

$$F_1(w) = \frac{D}{\sqrt{B}} [M(w) + N(w)]. \quad (G.5)$$

Defining

$$T_i(w) = G^{m_z+0.5} \int_{x_{min}^i}^{x_{max}^i} \frac{x^{m_z-0.5}}{\sqrt{1+x}} \exp(-\phi_z Gx) dx \quad (G.6)$$

and taking the integral limits as $\{x_{min}^1 = 0, x_{max}^1 = 1\}$ and $\{x_{min}^2 = 1, x_{max}^2 = P_{th}/G\}$ for $i = 1$ and $i = 2$, respectively, we have $M(w) = T_1(w)$ and $N(w) = T_2(w)$. Using [40, eq.3-a], and [38, 07.20.07.0001.01] for $i = 1$ and using [40, eq.3-b], and [32, eq. 3.351-2] for $i = 2$ in (G.6), $M(w)$ and $N(w)$ in (G.5) are calculated as

$$\begin{aligned} M(w) &= G^{m_z+0.5} \sum_{k=0}^{\Upsilon} (0.5)_k \frac{(-1)^k}{k!} \int_0^1 x^{m_z+k-0.5} e^{-\phi_z Gx} dx \\ &= G^{m_z+0.5} \sum_{k=0}^{\Upsilon} (0.5)_k \frac{(-1)^k}{k!} \frac{\Gamma(1)\Gamma(m_z+k+0.5)}{\Gamma(m_z+k+1.5)} \\ &\quad \times {}_1F_1(m_z+k+0.5; m_z+k+1.5; -\phi_z G) \end{aligned} \quad (G.7)$$

and

$$\begin{aligned}
N(w) &= G^{m_z+0.5} \sum_{j=0}^{\Upsilon} (0.5)_j \frac{(-1)^j}{j!} \int_1^{P_{th}/G} x^{m_z-1-j} \exp(-\phi_z Gx) dx \\
&= \sum_{j=0}^{\Upsilon} (0.5)_j \frac{(-1)^j}{j!} \frac{\Gamma(m_z - j, \phi_z G) - \Gamma(m_z - j, \phi_z P_{th})}{G^{-(j+0.5)} \phi_z^{(m_z-j)}}. \quad (G.8)
\end{aligned}$$

Finally, (G.5) is calculated by substituting (G.7) and (G.8). Then, (G.2) is calculated by replacing (G.3) and (G.5). As seen from (G.2), E_1 is independent of w whereas F_1 is a function of w .

APPENDIX H: Calculation of P_{s1}

Considering [80, eq. 13.1.12], [32, eq. 8.338-1], [32, eq. 8.331-1] and [80, eq. 6.1-22], (G.7) is simplified. Moreover, $G = B/C$, $B = 1/\bar{x}$ and $C = (1 + w\bar{y})/w\bar{y}$ are replaced in (G.7) and (G.8). After simplification, (G.5) is given as

$$F_1(w) = \frac{D}{\sqrt{B}} [M(w) + N_1(w) - N_2(w)], \quad (\text{H.1})$$

where

$$M(w) = \Lambda(k, n) \left(\frac{w\bar{y}}{1 + w\bar{y}} \right)^{n+m_z+0.5}, \quad (\text{H.2})$$

$$N_1(w) = \sum_{j=0}^{\Upsilon} (0.5)_j \frac{(-1)^j}{j!} \phi_z^{-(m_z-j)} \left(\frac{Bw\bar{y}}{1 + w\bar{y}} \right)^{j+0.5} \Gamma(m_z - j, \phi_z \frac{Bw\bar{y}}{1 + w\bar{y}}) \quad (\text{H.3})$$

and

$$N_2(w) = \beta(j) \left(\frac{w\bar{y}}{1 + w\bar{y}} \right)^{j+0.5}. \quad (\text{H.4})$$

In (H.2) and (H.4), we have

$$\Lambda(k, n) = \sum_{k=0}^{\Upsilon} \sum_{n=0}^{\Upsilon} \frac{(0.5)_k (-1)^k (-1)^n \phi_z^n}{k! n! (m_z + k + n + 0.5)} B^{n+m_z+0.5} \quad (\text{H.5})$$

and

$$\beta(j) = \sum_{j=0}^{\Upsilon} (0.5)_j \frac{(-1)^j B^{j+0.5} \Gamma(m_z - j, \phi_z P_{th})}{j! \phi_z^{(m_z-j)}}, \quad (\text{H.6})$$

where k , n and j are the indices of infinite summations defined in $\Lambda(k, n)$ and $\beta(j)$, respectively. Using [38, 06.06.26.0005.01], and [32, eq. 9.931-5], simplifying and substituting in (H.3), we have

$$N_1(w) = \alpha(j) G_{12}^{20} \left(\phi_z B \frac{w\bar{y}}{1 + w\bar{y}} \middle| j + 0.5, m_z + 0.5 \right), \quad (\text{H.7})$$

where

$$\alpha(j) = \left(\frac{1}{\phi_z} \right)^{m_z+0.5} \sum_{j=0}^{\Upsilon} (0.5)_j \frac{(-1)^j}{j!}. \quad (\text{H.8})$$

Finally, the simplified version of $F_1(w)$ in (H.1) is calculated by replacing (H.2), (H.7) and (H.4). Then, substituting (H.1) and (G.3) in (G.2), and, after some mathematical simplifications, P_{s1} in (3.16) is given as

$$P_{s1} = \int_0^{P_{th}} A(w) f_W(w) dw = \frac{a}{2} [E_1 E_2 - L], \quad (\text{H.9})$$

where

$$E_1 E_2 = DI \gamma(m_z, P_{th} \phi_z) \gamma(m_w, P_{th} \phi_w) / \phi_w^{m_w} \phi_z^{m_z} \quad (\text{H.10})$$

and

$$L = \frac{D}{\sqrt{B}} [L_1 + L_2 - L_3]. \quad (\text{H.11})$$

(H.10) is calculated by substituting (G.3) and (I.3). In (H.11), substituting (H.2), (H.7) and (H.4) and, after mathematical simplifications, we have

$$L_1 = \int_0^{P_{th}} M(w) f_W(w) dw = I \Lambda(k, n) \Delta_1, \quad (\text{H.12})$$

$$\begin{aligned} L_2 &= \int_0^{P_{th}} N_1(w) f_W(w) dw = I \alpha(j) \int_0^{P_{th}} w^{m_w-1} \exp(-w \phi_w) \\ &\times G_{12}^{20} \left(\phi_z B \frac{w \bar{y}}{1 + w \bar{y}} \middle| j + 0.5, m_z + 0.5 \right) dw \end{aligned} \quad (\text{H.13})$$

and

$$L_3 = \int_0^{P_{th}} N_2(w) f_W(w) dw = I \beta(j) \Delta_2. \quad (\text{H.14})$$

Please note that (H.13) is calculated numerically since a closed-form solution is not tractable. In (H.12) and (H.14), we have

$$\Delta_i = \int_0^{P_{th}} w^{m_w-1} \exp(-w \phi_w) \left(\frac{w \bar{y}}{1 + w \bar{y}} \right)^{u_i} dw, \quad (\text{H.15})$$

where for $i = \{1, 2\}$, we define $u_1 = n + m_z + 0.5$ and $u_2 = j + 0.5$, respectively, for (H.12) and (H.14). Taking $x = w \bar{y}$ and simplifying (H.15), (H.12) and (H.14) are rewritten as

$$L_1 = I \Lambda(k, n) \left(\frac{1}{\bar{y}} \right)^{m_w} (\lambda_1 + \varpi_1) \quad (\text{H.16})$$

and

$$L_3 = I \beta(j) \left(\frac{1}{\bar{y}} \right)^{m_w} (\lambda_2 + \varpi_2), \quad (\text{H.17})$$

respectively, where

$$\begin{aligned} \lambda_v &= \int_0^1 x^{m_w+u_v-1} (1+x)^{-u_v} \exp(-x \phi_w / \bar{y}) dx \\ &= \sum_{\delta=0}^{\infty} (u_v)_{\delta} \frac{(-1)^{\delta}}{\delta!} \int_0^1 x^{m_w+u_v+\delta-1} \exp(-x \phi_w / \bar{y}) dx \\ &= \sum_{\delta=0}^{\infty} (u_v)_{\delta} \frac{(-1)^{\delta}}{\delta!} \frac{1}{m_w + u_v + \delta} {}_1F_1 \left(m_w + u_v + \delta; m_w + u_v + \delta + 1; -\frac{\phi_w}{\bar{y}} \right). \end{aligned} \quad (\text{H.18})$$

(H.18) is calculated and simplified using [40, eq. 3-a], [80, eq. 13.2.1] and considering [32, eq. 8.338-1] and [32, eq. 8.331-1]. Please note that in (H.18), $v = 1$ and $v = 2$ for (H.16) and (H.17), respectively. Moreover, in (H.16) and (H.17),

$$\begin{aligned}\varpi_v &= \int_1^{P_{th}\bar{y}} x^{m_w+u_v-1} (1+x)^{-u_v} \exp(-x\phi_w/\bar{y}) dx \\ &= \sum_{\xi=0}^{\Upsilon} (u_v)_\xi \frac{(-1)^\xi}{\xi!} \left(\frac{\bar{y}}{\phi_w}\right)^{m_w-\xi} \left(\Gamma(m_w - \xi, \frac{\phi_w}{\bar{y}}) - \Gamma(m_w - \xi, \phi_w P_{th})\right).\end{aligned}\quad (\text{H.19})$$

(H.19) is calculated using [40, eq. 3-b] and [32, eq. 3.351-2]. (H.16) and (H.17) are calculated by substituting (H.18) and (H.19), respectively, for $v = 1$ and $v = 2$. (H.11) is calculated by substituting (H.16), (H.13) and (H.17). Finally, (H.9) is calculated by substituting (H.11) and (H.10). As seen from (H.9), P_{s1} is dependent on L , E_1 and E_2 .



APPENDIX I: Calculation of $A(z)$

In (3.19), substituting (3.14), $A(z)$ is written as

$$A(z) = \frac{a}{2} \int_0^{P_{th}} \left[1 - \frac{1}{\sqrt{C' + \frac{B'}{w}}} \right] f_W(w) dw, \quad (I.1)$$

where $C' = (1 + z\bar{y})/z\bar{y}$, $B' = 1/\bar{y}$, and $f_W(w)$ is given in (3.23). Noting that $A(z)$ given in (I.1) is a function with parameter z and applying the same procedure from (G.2) to (G.8), we have

$$A(z) = \frac{a}{2} (E_2 - F_2(z)), \quad (I.2)$$

where

$$E_2 = \int_0^{P_{th}} f_W(w) dw = I\gamma(m_w, P_{th}\phi_w)/\phi_w^{m_w} \quad (I.3)$$

and

$$F_2(z) = \frac{I}{\sqrt{B'}} [M'(z) + N'(z)]. \quad (I.4)$$

In (I.4),

$$\begin{aligned} M'(z) &= G_2^{m_w+0.5} \sum_{k=0}^{\Upsilon} (0.5)_k \frac{(-1)^k \Gamma(1) \Gamma(m_w + k + 0.5)}{k! \Gamma(m_w + k + 1.5)} \\ &\quad \times {}_1F_1(m_w + k + 0.5; m_w + k + 1.5; -\phi_w G_2) \end{aligned} \quad (I.5)$$

and

$$N'(z) = \sum_{j=0}^{\Upsilon} (0.5)_j \frac{(-1)^j (\Gamma(m_w - j, \phi_w G_2) - \Gamma(m_w - j, \phi_w P_{th}))}{j! G_2^{-(j+0.5)} \phi_w^{(m_w-j)}}, \quad (I.6)$$

where $G_2 = B'/C'$. Please note that E_2 is independent of z in (I.2).



APPENDIX J: Calculation of Average Powers

The average harvested powers at S and R are given as

$$P^L(\Xi, i, \rho) = \int_0^\infty \rho f_\rho(\rho) d\rho = \Xi \Gamma[m_\rho + 1] / \phi_\rho^{m_\rho + 1} \quad (\text{J.1})$$

and

$$P^{NL}(\Xi, i, \rho) = \int_0^{P_{th}} \rho f_\rho(\rho) d\rho + P_{th} B_i = \Xi \gamma(m_\rho + 1, P_{th} \phi_\rho) / \phi_\rho^{m_\rho + 1} + P_{th} B_i \quad (\text{J.2})$$

for linear and non-linear EH models, and are calculated by applying [32, 3.326-2] and [32, 3.351-1], respectively. Considering (J.1), the average powers at both S and R for the linear EH model are $P_s^L = P^L(D, 2, z)$ and $P_r^L = P^L(I, 1, w)$, while the average powers of S and R for the non-linear EH model are obtained considering (J.2) as $P_s^{NL} = P^L(D, 2, z)$ and $P_r^{NL} = P^L(I, 1, w)$, respectively. Here, $f_Z(z)$ and $f_W(w)$ are given in (3.21) and (3.23) and B_1 as well as B_2 are defined and calculated in (3.22) and (3.20), respectively.



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